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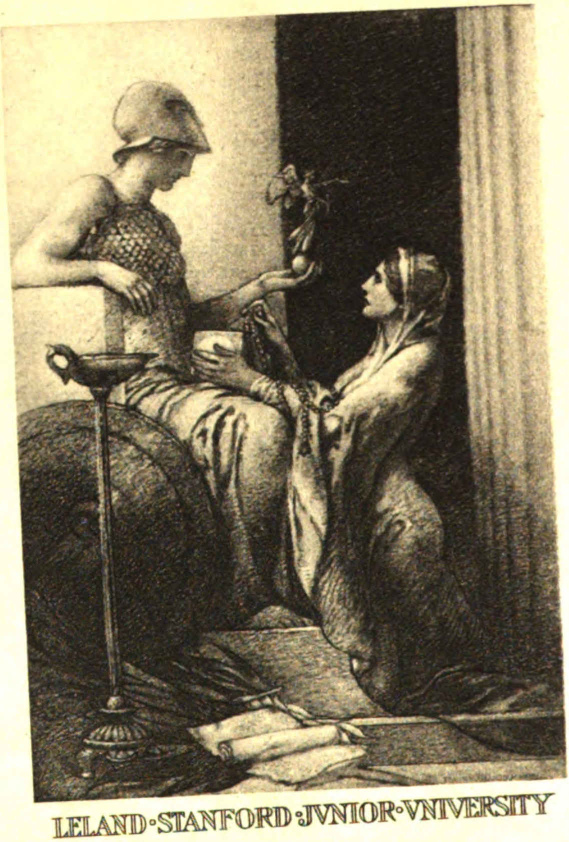

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VOLUME 8

1920

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PUBLISHED EVERY TWO MONTHS BY
THE INSTITUTE OF RADIO ENGINEERS
(INC.)

THE COLLEGE OF THE CITY OF NEW YORK

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NEW YORK, N. Y.

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NAVAL AIRCRAFT RADIO*

By

T. JOHNSON, JR.

(EXPERT RADIO AID, NAVY DEPARTMENT, WASHINGTON, D.C.)

INTRODUCTION

The importance of radio as applied to military aircraft is too self-evident to require extended comment. One of the most important functions of military aircraft is that of observing. In rapidly and accurately communicating at once the results of observation to a distant point lies the greatest value of obtaining such observations. Radio adds to this very great advantage that of being able to control the movements of aircraft from the ground or from other aircraft, and that of transmission of distress signals from disabled craft.

The naval aircraft radio problem is of a different character in many ways from that of the land military forces in that it introduces the use of this communication to a large extent in connection with anti-submarine and other coastal patrol duties where larger craft are used, and where larger and longer range radio sets are required. In these cases a specially trained operator is provided solely for radio, permitting the use of a set of greater complication of operation and greater transmission possibilities. This patrol duty involves the reporting of position as the aircraft covers its patrol territory, and the reporting of enemy craft or mines sighted, or vessels in distress. In connection with these duties there is involved that of convoy in which radio is useful in the aircraft communicating directly with the vessels under convoy.

The other phase of the naval aircraft problem in which radio plays a very important part is that of fire control from battleships. In this case the craft used are smaller, the radio usually being operated by the pilot, and the transmitting distance required is relatively short. Thus, from a radio viewpoint,

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* Received by the Editor, May 22, 1919. Presented before The Institute of Radio Engineers, New York, June 4, 1919; and before the Washington Section, June 24, 1919.

naval aircraft radio is divided into two distinctly separate phases, calling for apparatus and equipment of a widely varying character.

At the beginning of the war there was doubtless no field of radio work newer than that of aircraft radio. Like many of the other new technical problems introduced by war demands, that of aircraft radio was attended by multiple difficulties. The problem faced was that of providing large quantities of equipment which would be simple and inexpensive to manufacture, small, light weight, and possessing extreme ruggedness and simplicity of control. In meeting this problem there arose the development difficulties of providing new methods of investigation as applied to aircraft, and training personnel to conduct these investigations from a basic knowledge which was extremely meager. It was necessary to provide a large number of aircraft of the various standardized types for testing purposes; this was difficult, due to the general lack of aircraft at the beginning of the war. It was also discovered that in the development work it was necessary to employ pilots who were rather sympathetic with the radio investigations in order to obtain the most satisfactory results in the shortest time.

After the preliminary investigations had been conducted it was necessary for the apparatus to pass rapidly from the development to the standardization stage. In standardization it was necessary to combine smallness, lightness of weight, and manufacturing simplicity with simplicity of control, waterproofness, and the highest degree of ruggedness in ability to withstand shock of a widely varying nature. Standardization was also attended by the difficulty of simultaneous standardization of radio equipment with that of aircraft itself.

Installation difficulties were largely solved by the careful choice of complete equipments including all detailed fittings and material necessary for a standard installation. The installation work required, however, the special training of personnel who would be familiar with aircraft so that the installations might not impair the general utility of the craft. The initial installations were made in a standard manner by equipping each aircraft before it was shipped from the aircraft factory.

The matter of operation also involved the special training of personnel. Radio operating on aircraft is of a very special and unusual nature. The operator must usually work in a space which is more or less restricted, under a large number of conditions such as motor noise and rough flying which seriously distract his attention from his radio duties. The use of more recent forms of apparatus such as vacuum tube transmitters, regenerative receivers, and the radio compass, has still further necessitated special training.

EARLY DEVELOPMENT

The first operation of radio equipment on a seaplane was conducted by the Navy Department on July 26, 1912, at the Naval Academy at Annapolis, Maryland. Ensign (now Lieutenant-Commander) Charles H. Maddox, U. S. N., acting as operator on the seaplane, transmitted a message from a height of 300 feet (91.5 meters) to the torpedo boat U. S. S. *Stringham*. The communication was maintained at that time up to a distance of three nautical miles (5.55 kilometers). The complete receiving apparatus was suspended in front of the operator by a strap passing over his shoulders as shown in Figure 1. A specially constructed helmet was used for holding the telephone receivers and keeping out external noises. A "balanced antenna" with similar portions in the upper and lower planes was employed.

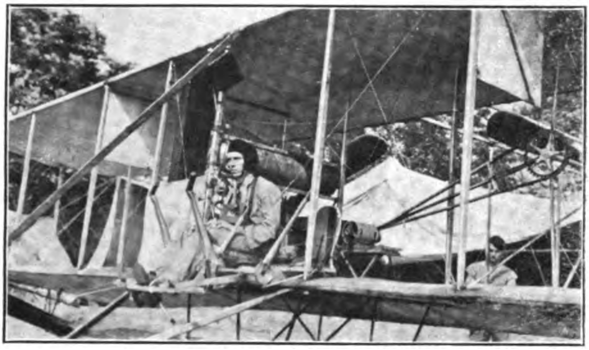


FIGURE 1

Early in 1916 Lieutenant-Commander (now Commander) S. C. Hooper, U. S. N., of the Radio Division, Bureau of Steam Engineering, Navy Department, realizing the growing importance of the application of radio to aircraft, urged the establishment of a laboratory for this work. Such a laboratory was started at the United States Naval Air Station, Pensacola, Florida.

The first apparatus tested was a transmitter, a 250-watt tuned vibrator set manufactured by William Dubilier. Three sets of this type were installed on Curtiss pusher seaplanes of the type illustrated in Figure 2. The transmitter was strapped to the inclined struts near the pilot's seats, power being supplied from a direct current generator geared to the camshaft of the engine. Transmission was conducted on a wave length of 300 meters, a single trailing wire antenna 150 feet (45.7 meters) in length being employed. The entire equipment installed weighed 80 pounds (36.2 kilograms). On May 15, 1916, Chief Electrician (now Chief Warrant Officer) S. S. Halliburton, U. S. N., transmitted the first official radio message from a United States Naval seaplane in flight a distance of 20 nautical miles (37.1 kilometers) to the cruiser U. S. S. *North Carolina*.

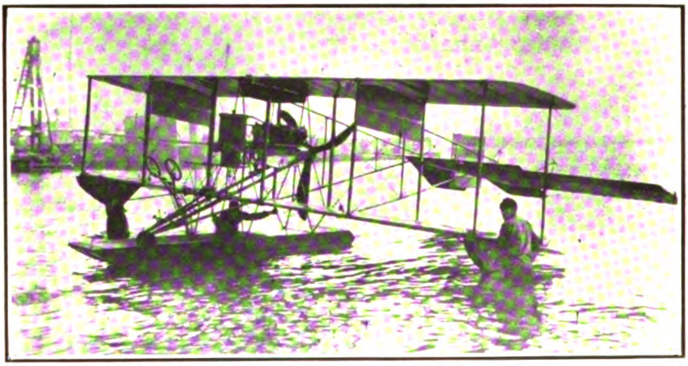


FIGURE 2

The results of these tests so proved to the Navy Department the importance of providing aircraft with radio equipment that it was decided to purchase seventy-five seaplane transmitting sets at once. From the observations made in the preliminary investigations at Pensacola specifications were drawn up, the use of a propeller-driven generator and a single trailing wire antenna being called for, the metallic parts of the seaplane to be used as counterpoise. A total weight of 100 pounds (45.4 kilograms) was allowed for the equipment. Of the large number of samples submitted by manufacturers three types were purchased, a spark set, a buzzer excitation set, and a vacuum tube transmitter. The spark set was type CM 295, manufactured by the Marconi Wireless Telegraph Company. The vibrator arc set was type CS 350, invented by William Dubilier and manufactured by the Sperry Gyroscope Company and illustrated in Figure 3. The vacuum tube set, type CF 118, manufactured by the De Forest Radio Telephone and Telegraph Company and illustrated in Figure 4, was the first radio telephone equipment to be installed on aircraft.

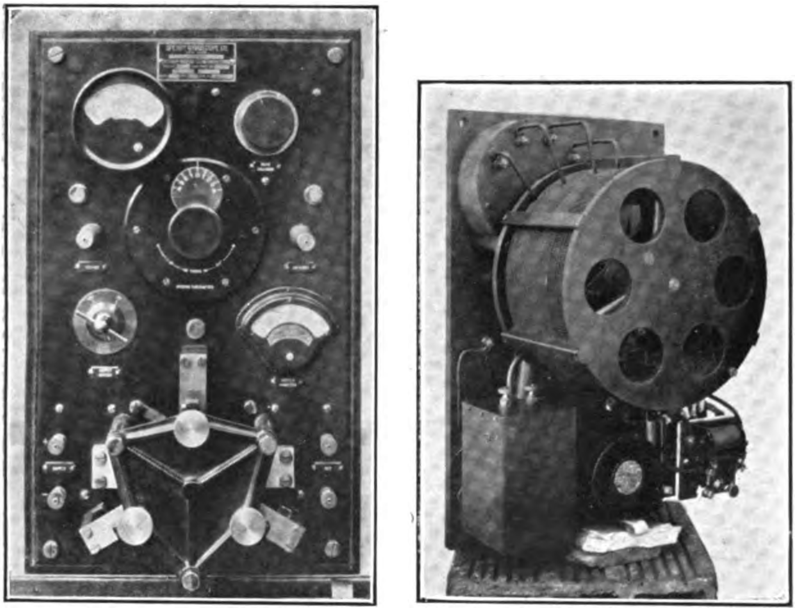


FIGURE 3

In the spring of 1917, the personnel at the station at Pensacola was considerably increased and more extended work in aircraft radio pursued. Measurements were made of antenna constants on seaplanes and the directive effect of trailing wire antennas was investigated. It became apparent that a satisfactory intercommunicating system between pilot and radio operator would be essential to the most effective use of the radio apparatus and a very satisfactory equipment of the voice tube type was developed together with a suitable helmet and other appurtenances.

At this time the Navy Department also realized the importance of the radio compass for aircraft, and investigations were started upon this subject. Experimental work was also conducted on the use of a high tension ignition magneto as a trans-

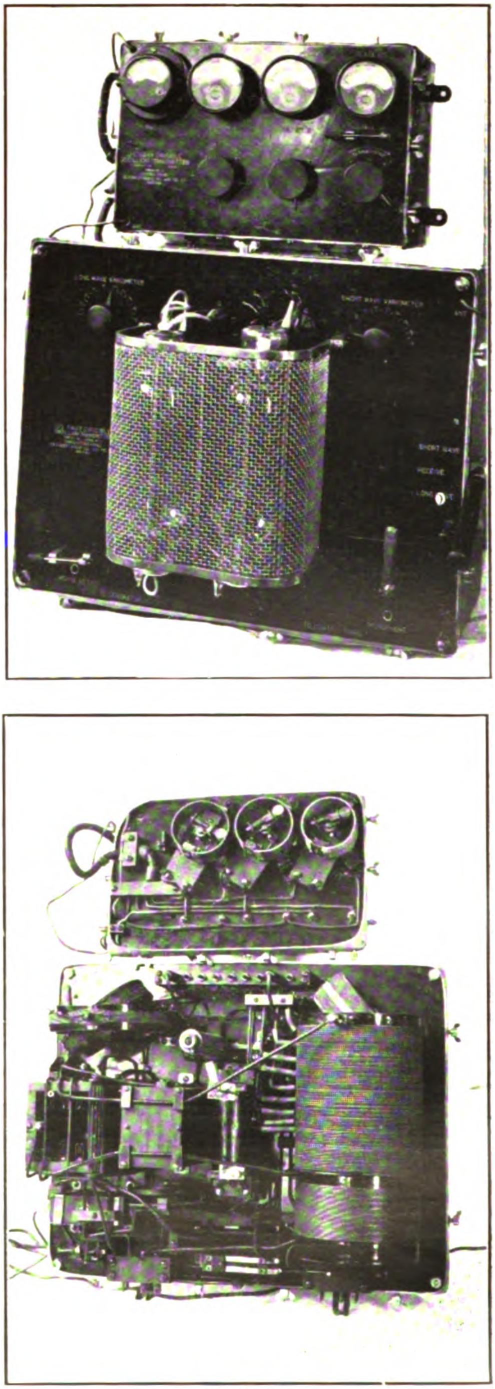

LONG WAVE VARIOMETER
SHORT WAVE VARIOMETER
ANT
SHORT WAVE
RECEIVE
MICROPHONE

FIGURE 4

mitter of radio signals including the design of a key suitable for handling the high voltages involved. Development of installation fittings such as antenna reels and antenna weights was also undertaken.

At this time there was developed for the Navy Department the 0.5-kilowatt quenched spark aircraft transmitter, type CE 615, manufactured by E. J. Simon, New York, and illustrated in Figure 5. This transmitter was a very great step in advance over any radio transmitters which had been previously designed for aircraft use. The transmitter consisted essentially of a panel on which were mounted the main elements of a quenched spark set, the top of the panel being rounded to fit the standard airplane fuselage in use at that time. A reel was supplied with this set which was made entirely of insulated material so that tuning of the antenna circuit could be accomplished by the variation of the length of trailing antenna while the transmitter was operated. Power was supplied from a propeller-driven generator mounted on the wing of the airplane, a brake being provided to prevent the propeller from revolving when the radio set was not in use. This set completely installed weighed

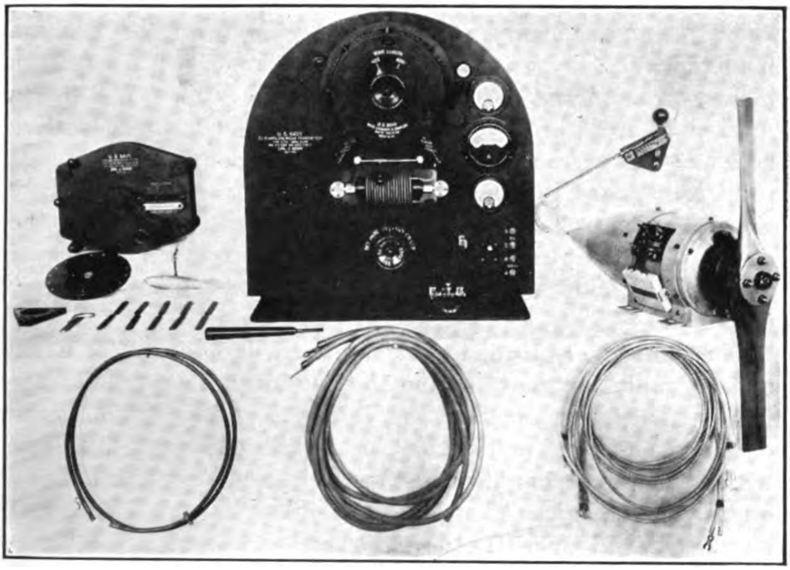


FIGURE 5

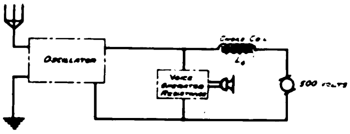
approximately 100 pounds (45.4 kilograms). During the summer of 1917, signals were transmitted a distance of 150 nautical miles (278 kilometers), using this transmitter.

The only aircraft receiver of any practical value developed during this period was Type CE 937, manufactured by E. J. Simon, New York. This receiver contained in a case 13 inches (33 cm.) wide by 7 inches (17.8 cm.) high by 7.25 inches (18.4 cm.) deep and weighing 10.5 pounds (4.77 kilograms), was of the single circuit type having a wave length range of approximately 275 to 2,500 meters. A single variable air condenser was provided for tuning, being connected in series with the antenna inductance for low wave lengths and in parallel with it for the higher wave lengths. A single vacuum tube Type CW 933 was utilized as a detector, provision being made for regenerative action. The vacuum tube was carefully mounted on sponge rubber to guard it from shock. The receiver was extremely simple to control and was a distinct accomplishment in the design of aircraft radio apparatus at that time. The lack of additional amplification, however, soon rendered this receiver obsolete.

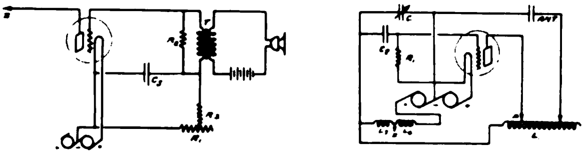
On January 1, 1918, the experimental laboratory for naval aircraft radio was moved from the station at Pensacola to the Naval Air Station at Hampton Roads, Norfolk, Virginia, and development work undertaken on a far more extensive scale with a view to accomplishing standardization of equipment and quantity production as soon as possible. At this time great stress was laid upon the development of vacuum tube transmitters for telephone use, and the radio direction finder apparatus. At this time there also became available flying boats of the latest standardized type, thereby permitting the standardization of radio installations.

The preliminary experimental work at Hampton Roads involved a very large number of fundamental investigations in connection with various details such as all forms of power generating apparatus including propellers, storage batteries, generators and dynamotors, all forms of antenna and ground systems, electrical intercommunicating systems, helmets, microphones, and the many other units forming a part of complete equipments.

The first tests of radio transmitters conducted at Hampton Roads were those made upon a vacuum tube transmitter developed by the Western Electric Company. This was in the form of a preliminary model utilizing ten 5-watt tubes, type CW 931, five being used as oscillators and five as modulators. The power was supplied from a 12-volt storage battery, the filaments being lighted directly from this battery, and the power for the plate circuits, 0.6 ampere at 300 volts, being supplied from a dynamotor run from the storage battery. The set was arranged for telephone or buzzer modulated telegraph transmission on 800 or 1,600 meters, antenna currents of 1.5 amperes and 1 ampere respectively, being obtained on these wave lengths. The weight of the transmitting and receiving set, including storage battery sufficient for 45 minutes' intermittent operation, was 185 pounds (91.3 kilograms). The modulator system used was the Heising system, the fundamental principle of which is indicated in Figure 6. The Colpitts' oscillator system, also illustrated in Figure 6, was used. The receiving set was the vacuum tube type including a detector and three stages of audio frequency amplification. The best transmitting distances obtained from aircraft with this set were 45 nautical miles (83.4 kilometers) by telephone and 100 nautical miles (185 kilometers) by buzzer modulated telegraph on a wave length of 600 meters.



HEISING MODULATION SYSTEM



VOICE OPERATED RESISTANCE OR MODULATING TUBE.

COLPITTS OSCILLATION CIRCUIT.

FIGURE 6

The second set developed by the Western Electric Company and tested at Hampton Roads was that shown in Figure 7. This set was arranged for transmission on 1,540 meters on a trailing wire antenna and on 600 meters on an umbrella type of antenna used when the seaplane was at rest on the water. Provision was made for continuous wave telegraph and telephone signals on both wave lengths. For telegraphing, eight type

CW 931 5-watt vacuum tubes were used operating in parallel, all as oscillators. The signals were made by operating a telegraph key which opened and closed the circuit supplying the plates of the oscillators. For telephoning, three of the tubes were used operating in parallel as oscillators, and the other five operating in parallel as modulators. Power for the set was supplied from a 24-volt storage battery from which the filaments were lighted directly, the plate circuits of the tubes being supplied from a 350-volt dynamotor which operated from the 24-volt storage battery. The schematic circuit diagram of this set is shown in Figure 8.

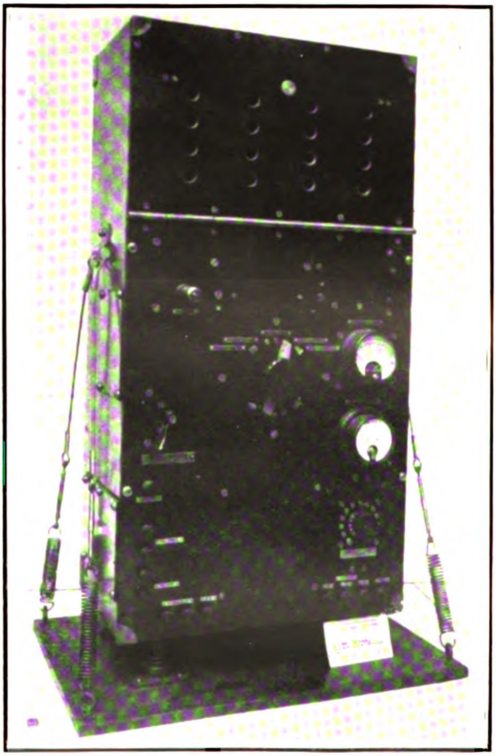


FIGURE 7

Operating this set on a trailing antenna, an antenna current of 1.6 amperes was obtained while telegraphing, transmission being conducted over a distance of about 100 nautical miles

(185 kilometers). The antenna current while telephoning was about 1 ampere, a range of ten nautical miles (18.5 kilometers) being the maximum over which satisfactory telephone communication was maintained. This short range was due to the improper functioning of the modulator, a defect which was not corrected as the development work on this set was discontinued due to its excessive size and weight.

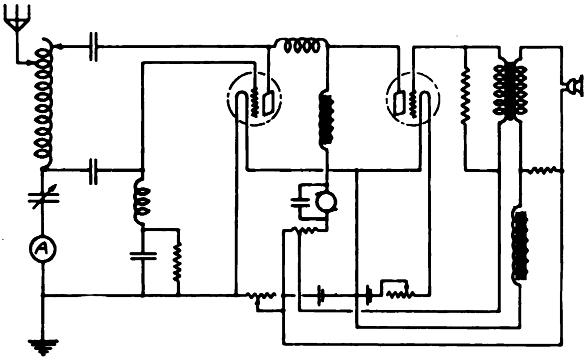


FIGURE 8—Western Electric Co. 100-mile set, Telephone Connection

Another set tested at Hampton Roads was one designed by the De Forest Radio Telephone and Telegraph Company, shown in Figure 9, and employing two De Forest tubes. It was arranged for transmission over two ranges of wave length, from 350 to 600 meters and from 1,000 to 2,000 meters, telephone and "chopped" continuous wave telegraph signals being used. Power for the plate circuits was supplied at 1,200 to 1,500 volts direct current from a propeller-driven generator. The filaments of the tubes were lighted from a 12-volt storage battery from which was also driven a small motor which operated the continuous wave interruptor or "chopper." The schematic circuit is shown in figure 10. Telegraph signals were transmitted for a distacne of 125 nautical miles (232 kilometers) from seaplane to shore but, no satisfactory telephone operation of this set was obtained.

There was also tested at Hampton Roads an 0.5-kilowatt quenched spark transmitter of improved type designed by E. J. Simon and illustrated in Figure 11. This transmitter was very compact and light, the total weight of the complete equipment being 42 pounds (18.1 kilograms). The transmitter in which the antenna reel was mounted was 12 by 11 by 11 inches (30.48

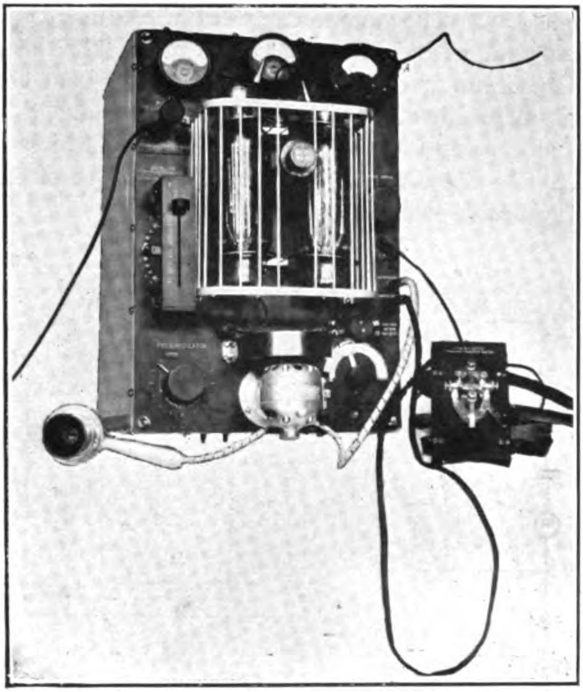


FIGURE 9

by 27.9 by 27.9 centimeters). Transmission on a single wave length of 375 meters was arranged for, power being obtained from a propeller-driven generator. The inductances used were

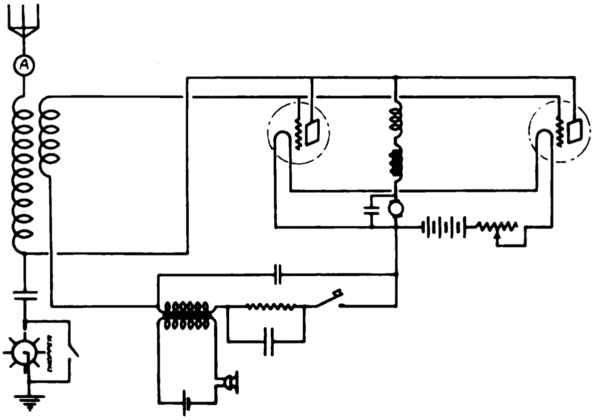
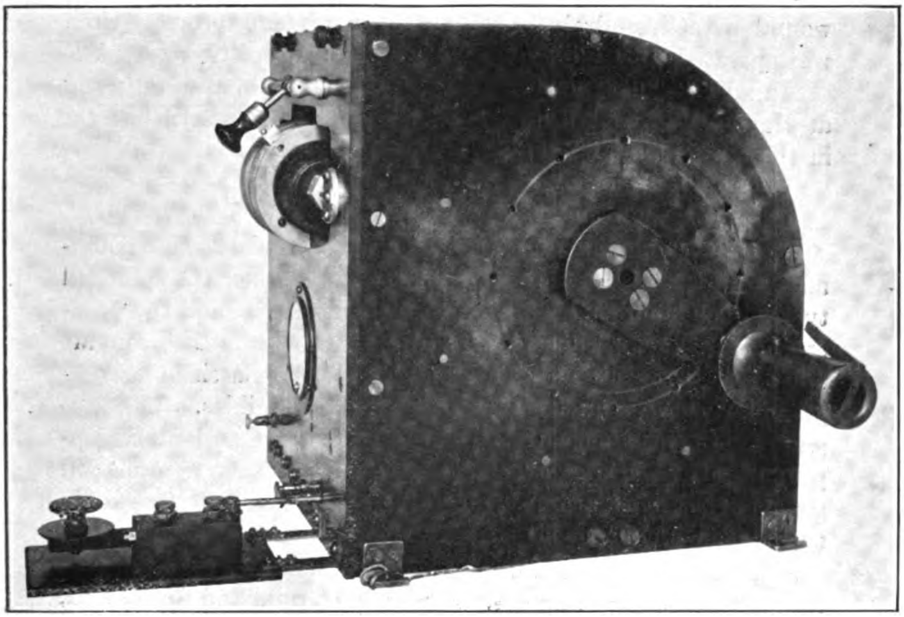


FIGURE 10



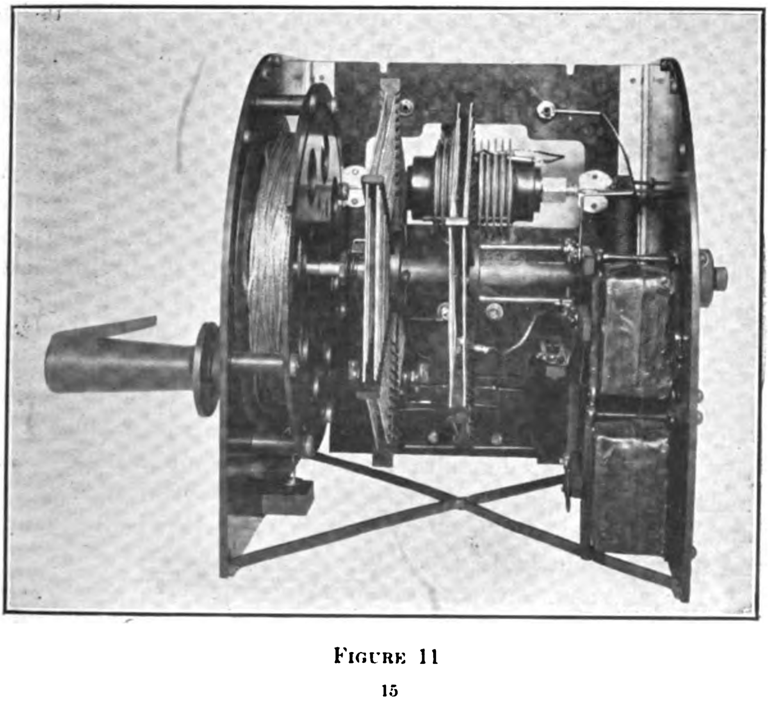


FIGURE 11

wound with litzendraht. A novel form of send-receive switch was operated by hinged movement of the transmitting key base.

In addition to the above experimental sets there were tested at Hampton Roads all of the standardized sets described later in this paper.

STANDARD NAVAL AIRCRAFT AND ANTENNA CONSTANTS

In reviewing the problem of providing standard radio equipment for naval aircraft, it is of interest to consider the standard types of craft used, especially as regards their size and the properties of the radiating systems applied to them. The aircraft used in the naval service may be generally classified as seaplanes, ship airplanes, flying boats, and dirigibles. The seaplane consists essentially of an airplane fuselage and wings supported by means of struts on pontoons. Seaplanes are used generally for training purposes only and consist of two main types known as N-9 and R-6.

Ship airplanes are small single-seat airplanes used essentially for fire control duty and designed for flight from and to the deck of a battleship.

Flying boats consist of a light boat body which accommodates passengers and fuel tanks and to which are mounted wings and tail stabilizers. The engines are usually mounted in struts between the wings. Flying boats are used for patrol and other heavy service duties where landings must occasionally be made in fairly rough water. The four standard types of naval flying boat are known as HS-2-L, H-16-A, F-5-L, and NC.

The dirigibles used in the naval service are all of the non-rigid type and consist of several classes of varying size. The fundamental physical characteristics of the standard naval aircraft are outlined in the tabulation of Figure 12, and the craft are shown in the photographs following.

In the design of radio equipment it is, of course, of prime importance to determine the constants of the antennas used in order to properly standardize on antennas for the various sets. The measurement of antenna constants on aircraft is beset with the most unusual difficulties. Adjustments of various kinds must be made under conditions of severe mechanical vibration and external noise which render ordinary methods of measurements very difficult to use.

The antenna constants discussed herein were measured by employing a calibrated oscillator utilizing a 50-watt vacuum tube, type CG 1144, using a plate potential of 350 volts. With very

Type of Craft	Span		Chord		Gap		Overall Length		Overall Height		Weight Loaded		Useful Load		Crew	Maximum Speed*		Cruising Radius in Hours	
	Feet	Meters	Feet	Meters	Feet	Meters	Feet	Meters	Feet	Meters	Lbs.	Kilograms	Lbs.	Kilograms		Naut. miles per hour	Kilometers per hour	At full speed	At cruising speed
Seaplane N-9........	53	16	5	2	5	2	30	9	11	3	2,600	1,180	490	222	2	80	148	1.5	
Seaplane R-6........	57	17	6	2	7	2	33	10	14	4	3,945	1,790	895	406	2	82	152	4.2	
Ship Airplane.......	28	9	6	2	Monoplane		22	7	8	2	650	295	284	129	1	65	120	2	
Flying Boat HS-2-L..	74	23	6	2	8	2	38	12	15	5	6,432	2,910	2,135	966	3	91	168	4.4	6.4
Flying Boat H-16-A..	95	26	7	2	8	2	46	14	18	6	10,900	4,950	3,500	1,590	5	95	176	3.9	5.5
Flying Boat F-5-L....	104	32	8	2	9	3	49	15	19	6	13,000	5,900	4,750	2,160	5	83	154	5.5	7.9
Flying Boat NC......	126	38	12	3.7	14	4	68	21	24	7	22,100	10,040	8,000	3,630	6	83	154	9.0	13.0

*Cruising speed is 66 per cent. of Maximum Speed.

Class of Dirigibles	Envelope						Length of Car		Useful Load		Crew	Speed				Cruising Radius			
	Length		Maximum Diameter		Capacity							Maximum		Cruising		At Cruising Speed with Normal Fuel		At Maximum Speed with Maximum Fuel	
	Feet	Meters	Feet	Meters	Cubic Feet	Cubic Meters	Feet	Meters	Lbs.	Kilograms		Naut. miles per hour	Kilometers per hour	Naut. miles per hour	Kilometers per hour	Naut. miles	Kilometers	Naut. miles	Kilometers
"B"	160	49	32	10	84,000	2,300	28	9	2,000	908	3	45	83	35	65	500	925	1,100	2,040
"C"	192	58	42	13	171,000	4,680	40	12	3,930	1,735	5–8	60	111	35	65	530	980	1,800	3,330
"D"	198	60	50	15	186,000	5,090	29	9	4,109	1,870	5–8	59	109	35	65	530	980	1,800	3,330
"E"	162	49	34	10	95,000	2,600			2,135	970	3	55	102	35	65	375	694	700	1,290
"G"	285	87	57	17	415,000	11,350	52	16	11,000	4,990	10	60	111	40	74	1,440	2,660	3,780	7,000

FIGURE 12—Approximate Data on size of Standard Naval Aircraft

loose coupling an antenna current of 125 milliamperes was thus easily obtained, and practically all the measurements were taken with a current of at least 100 milliamperes, a long scale instrument being used to assist in obtaining accurate results. With no added resistance in the antenna circuit and a meter reading of 100 milliamperes adjusted to by variation of coupling, resistances of 10 ohms and 20 ohms were successively inserted in circuit and current readings noted. The antenna resistance obtained in each case was compared, and, if results did not check, the set was again resonated and readings repeated.

The fundamental wave length was measured by coupling two small turns of the antenna circuit with the oscillator and resonating the circuit. To obtain the necessary range of wave lengths, coils of various inductance values were successively inserted. In computing the resistance of the antenna, the resistance of these coils at the frequencies used was deducted as well as the resistance of the meter, the set and the connecting wires forming the complete installation.

Great care was taken in arranging the circuit loading coils so that minimum capacity effects were experienced, and capacity currents which would distort the resistance results were avoided. The milliammeter was mounted on rubber cushions to absorb vibration. In spite of this it was often necessary to wait some time before the pitching and tossing of the aircraft decreased to a point where readings could be taken. In making these measurements the greatest of care was observed in so waiting until stable readings could be obtained.

In the results of these measurements, *effective capacity* is defined as that capacity which, used in conjunction with the antenna inductance, will resonate at the natural wave length of the antenna; and *equivalent capacity* at any loaded wave length as that capacity which, if used with an inductance equal to the loading inductance, will resonate at the same wave length as if the loading inductance were applied to the antenna.

Figure 13 illustrates the measuring apparatus set up in an H-16-A flying boat. The arrangement for obtaining variable coupling is shown on the left side against the ribs of the hull. The calibrated oscillator itself is shown in the foreground with the 125-milliampere meter mounted on a rubber suspension slightly above and to the right. The various loading coils used are on the shelf in the background.

The smallest type aircraft used in the naval service is the ship airplane illustrated in Figure 14, or several other types of similar

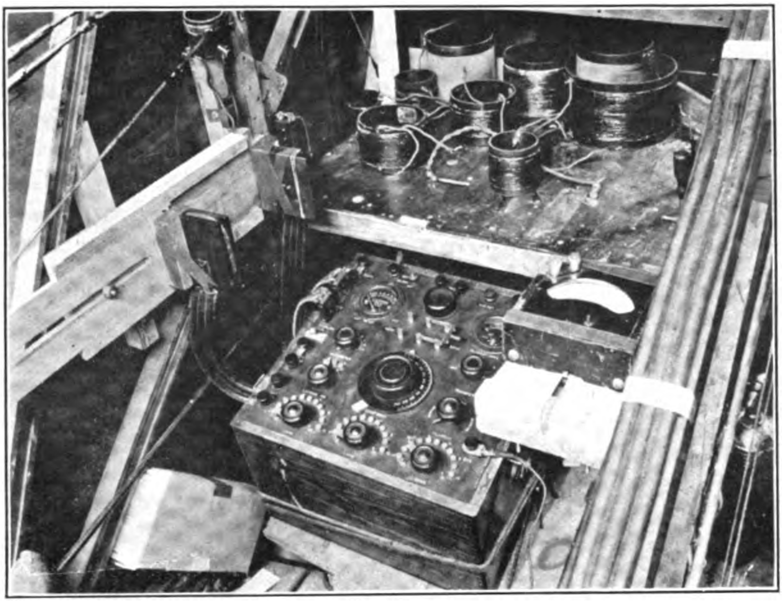


FIGURE 13

small craft designed for carrying a pilot only and for operation from the deck of a ship rather than from the water. With this type of craft there is used either a short single wire trailing antenna or the radiating system consists of wire networks in the two wings.

Figure 15 shows the type N-9 seaplane on a handling truck. The only type of antenna used on this craft is a single trailing wire, the constants of which are shown in Figures 16, 17, and 18.

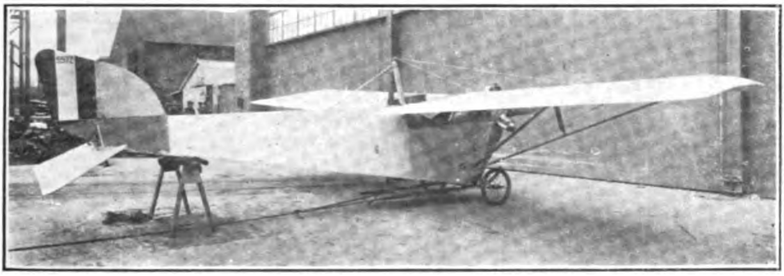


FIGURE 14

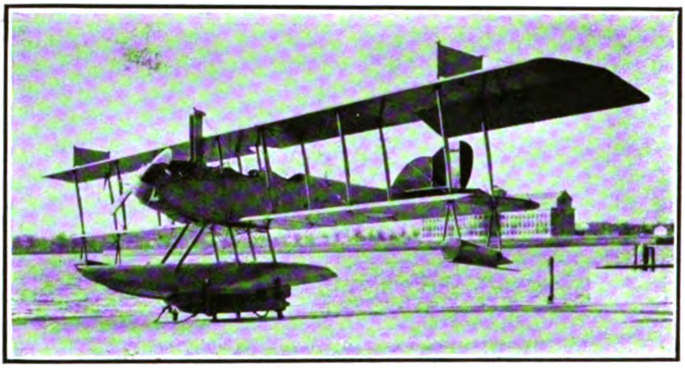
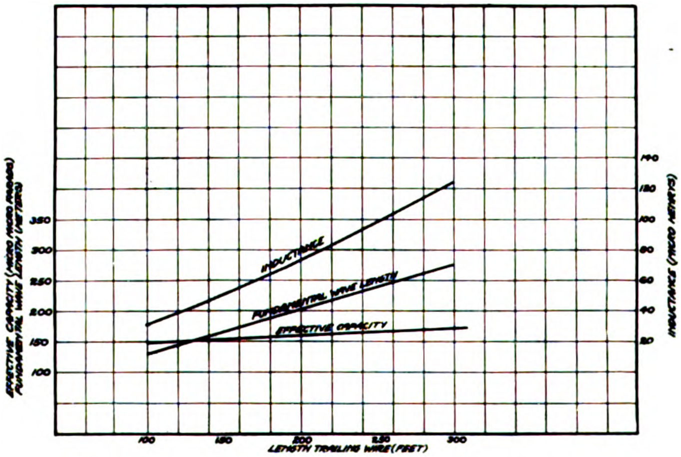
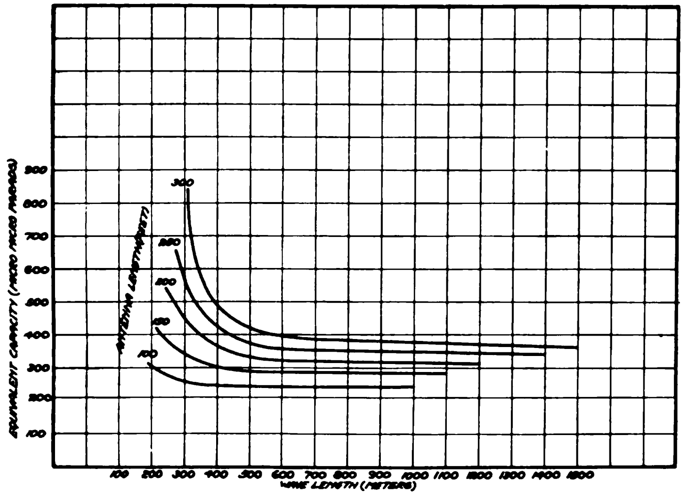


FIGURE 15



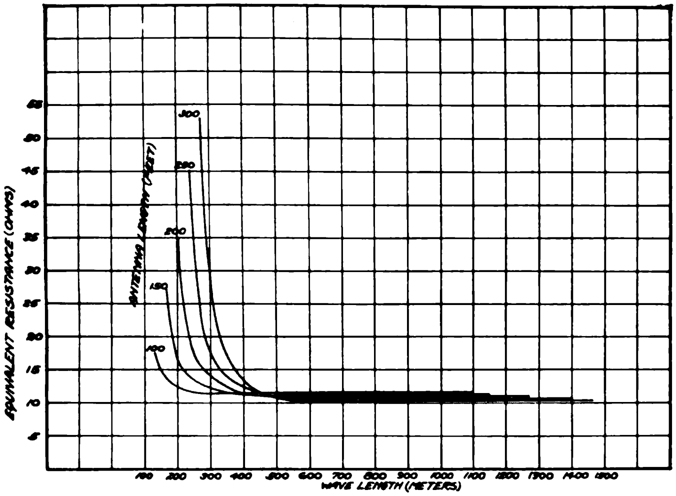
EFFECTIVE CAPACITY, INDUCTANCE
AND FUNDAMENTAL WAVE LENGTH
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
SEAPLANE TYPE N-9

FIGURE 16



EQUIVALENT CAPACITY
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
SEAPLANE TYPE N-9.

FIGURE 17



EQUIVALENT RESISTANCE
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
SEAPLANE TYPE N-9

FIGURE 18

Figure 19 shows a type HS-2-L flying boat. This boat is driven by a single 12-cylinder Liberty motor, the propeller being arranged as a pusher. The antenna systems on this craft are illustrated in Figure 20. The antenna used when in flight consists of a single trailing wire leading from the radio apparatus in the forward part of the boat thru a lead-out fixture and thru an insulator mounted on the under side of the lower port wing. The wire of standard length for the particular equipment used is wound on a metal reel by which the wire may be wound out

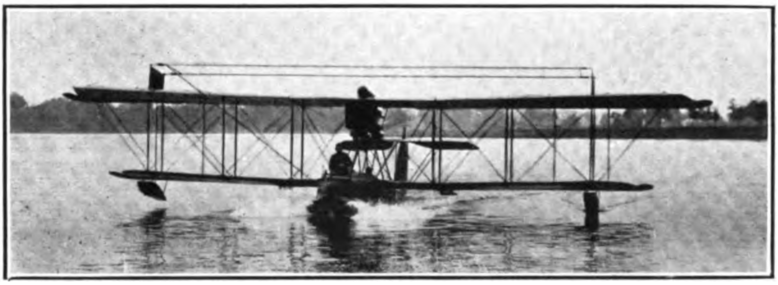


FIGURE 19

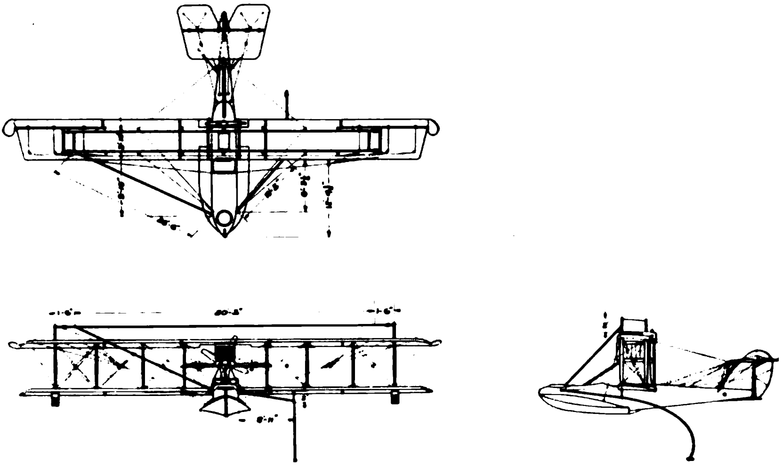


FIGURE 20

and in. The standard reel and fittings are illustrated in Figure 21. When in flight, the entire standard length of wire is unwound from the reel. The inner end of the wire is connected to the reel by a piece of insulating twine so that no high potentials are developed on the reel when the radio set is in operation An electrical connection is made to the trailing antenna wire thru the metal portion of the lead-out insulator. The standard

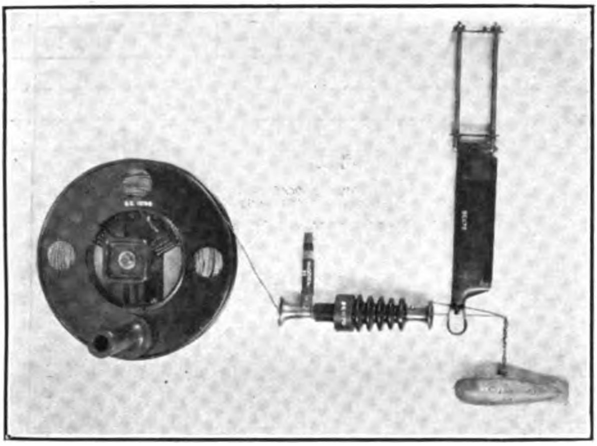
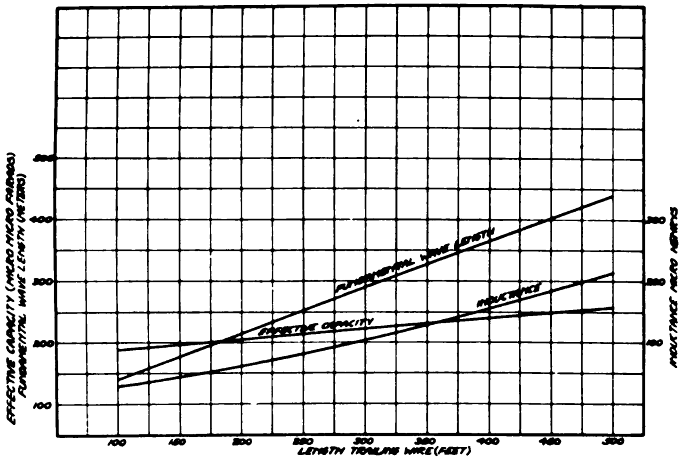


FIGURE 21

antenna wire used consists of six strands of seven silicon bronze wires of 0.008 inch (0.203 millimeter) cabled with strong 5-ply cotton thread. A two pound (0.91 kilogram) stream-line lead weight is attached by a swivel to the outer end of the trailing antenna. The International Radio Telegraph Company rendered invaluable assistance and co-operation in the development of the above reel.

The constants of this single wire trailing antenna are shown in Figures 22, 23, and 24. The counterpoise used consists of all metal pieces of the craft such as brace wires, flying wires, control wires, engines, and fuel tanks, electrically connected. These wires and fittings extend thru the entire wing area and body of the boat.

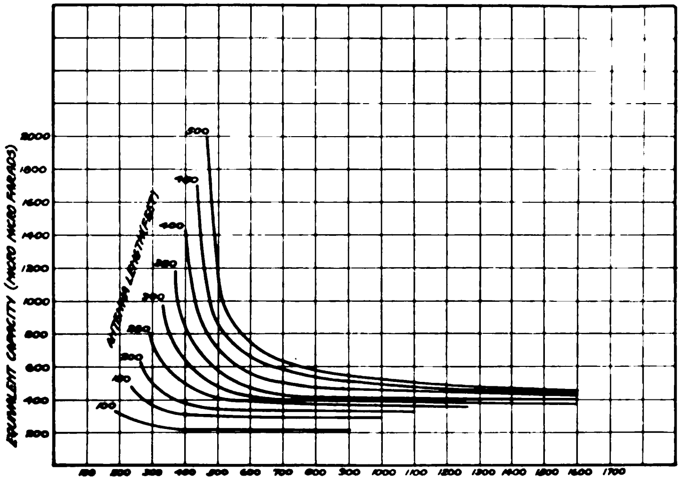
Another type of antenna used on this flying boat is known as the skid fin antenna and is mounted as a rectangle on the skid fins on the upper plane, the lead-in wire being carried from the forward starboard corner of the rectangle to a lead-in insulator on



EFFECTIVE CAPACITY, INDUCTANCE
AND FUNDAMENTAL WAVE LENGTH
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE HS-2-L

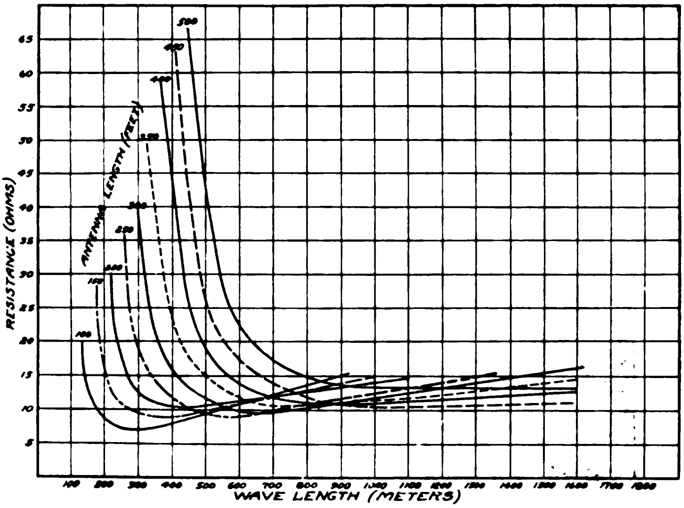
FIGURE 22

the starboard side of the boat forward, at a point corresponding to that at which the trailing antenna leads out on the port side.



EQUIVALENT CAPACITY
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE HS-2-L

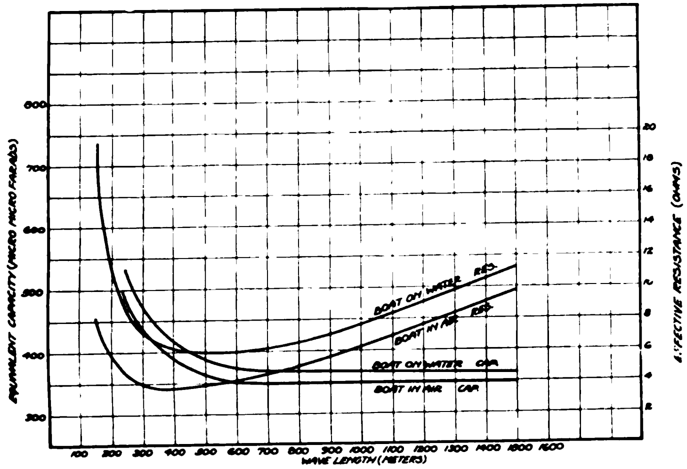
FIGURE 23



ANTENNA RESISTANCE
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE HS-2-L

FIGURE 24

This antenna must be used when the boat is resting on the water, but may also be used successfully when in flight, altho transmission distances obtained will not equal those obtained with the trailing antenna. The constants of this antenna, both when the boat is in flight and on the water, are shown in Figure 25.



EQUIVALENT CAPACITY & EFFECTIVE RESISTANCE
OF
SKID FIN ANTENNA
ON
FLYING BOAT TYPE HS-2-L

FIGURE 25

Figure 26 illustrates a type H-16-A flying boat. This boat is propelled by two 12-cylinder Liberty motors mounted outboard, the propellers usually being arranged to act as tractors. The

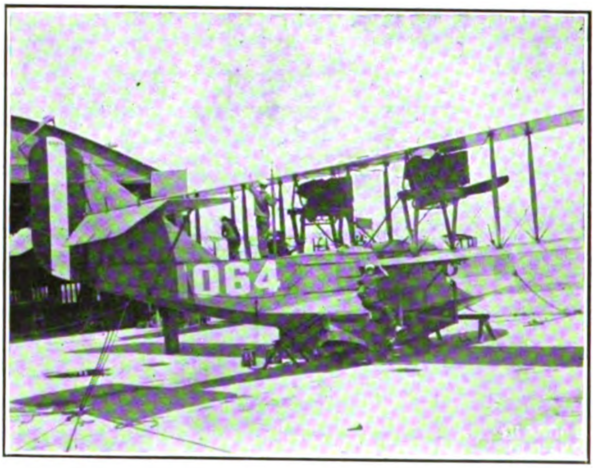


Figure 26

antenna systems employed are illustrated in Figure 27. Trailing wire antennas are also used on this type of boat with similar fittings to those used on the HS-2-L type, the main difference

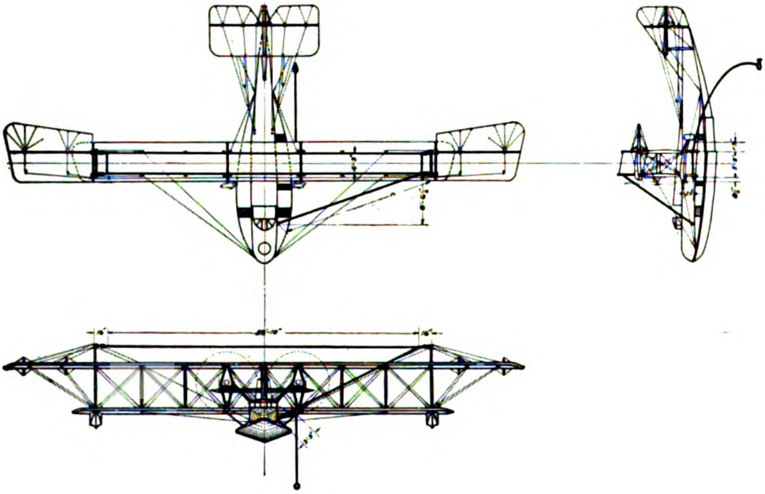
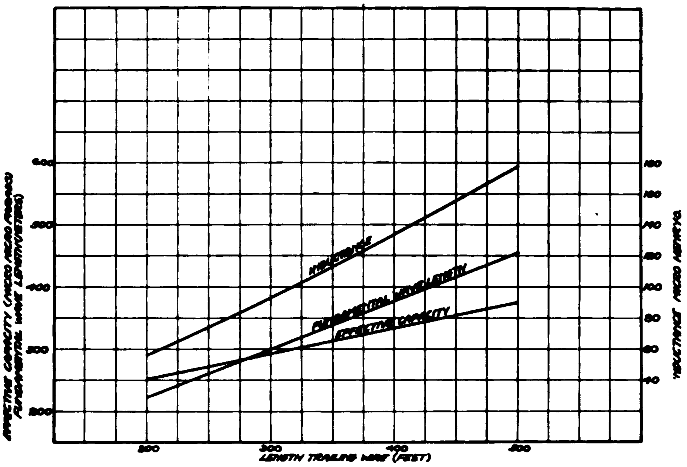


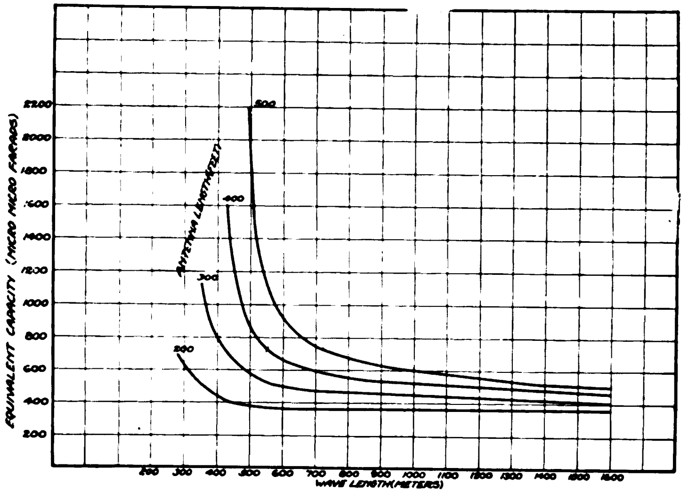
Figure 27

being that the radio apparatus is located just forward of the wings, and the antenna wires are led out from the hull at that point. In the case of this larger boat there has been used a double trailing antenna, the two wires being led off symmetrically on the two sides of the boat. The antenna characteristics of both single and two-wire trailing antennas are illustrated in Figures 28, 29, 30, 31, 32, and 33.



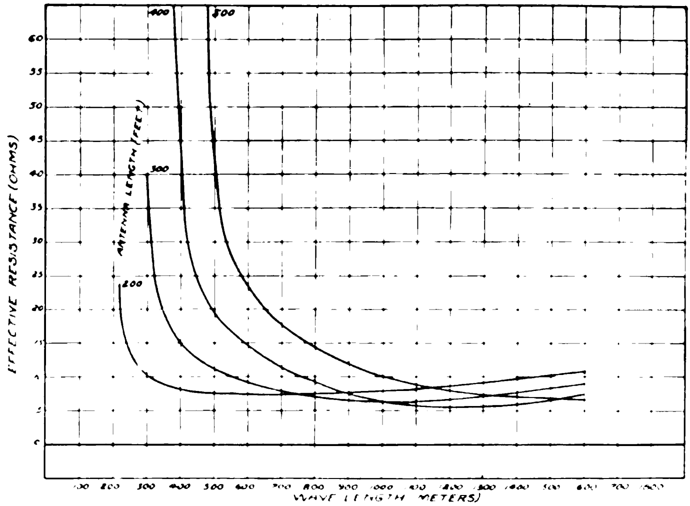
EFFECTIVE CAPACITY, INDUCTANCE
AND FUNDAMENTAL WAVE LENGTH
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

Figure 28



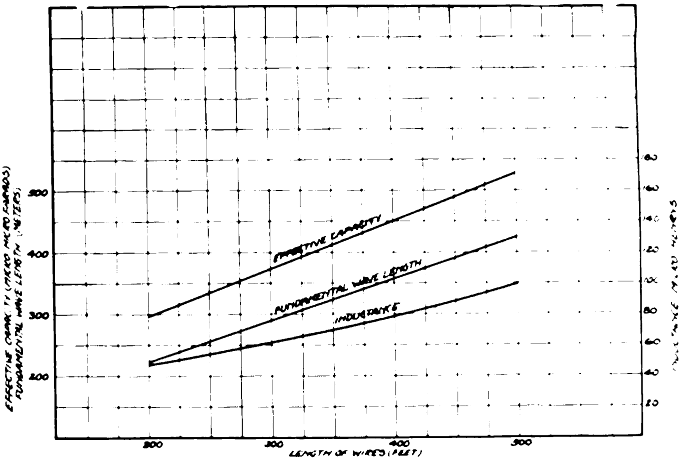
EQUIVALENT CAPACITY
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

Figure 29



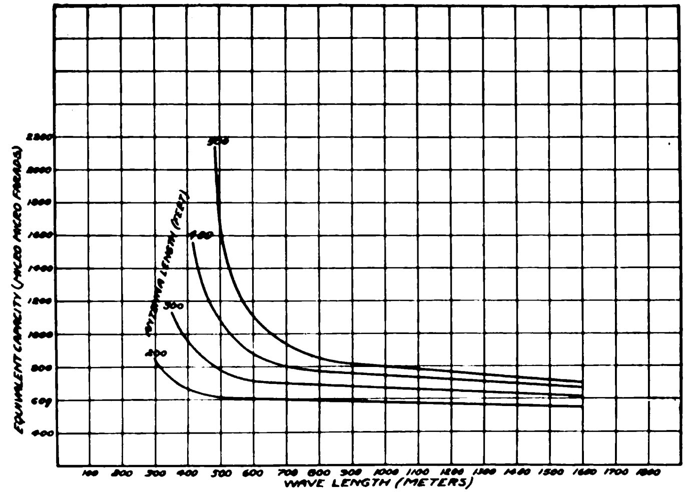
EFFECTIVE RESISTANCE
OF
STANDARD SINGLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

FIGURE 30



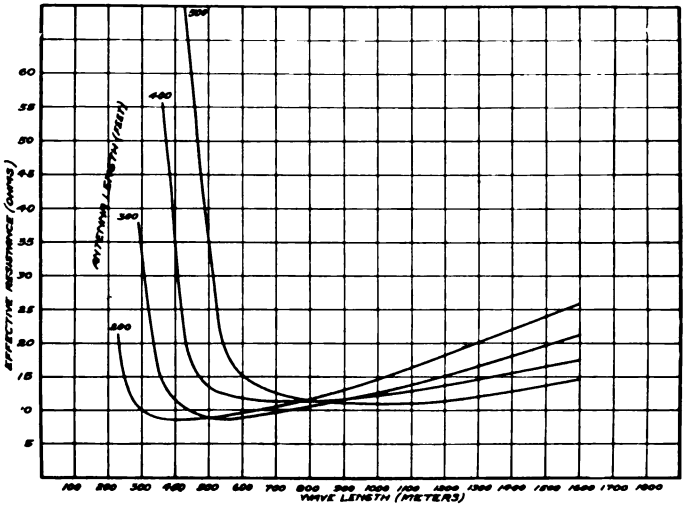
EFFECTIVE CAPACITY INDUCTANCE
AND FUNDAMENTAL WAVE LENGTH
OF
DOUBLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

FIGURE 31



EQUIVALENT CAPACITY
OF
DOUBLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

FIGURE 32



EFFECTIVE RESISTANCE
OF
DOUBLE WIRE TRAILING ANTENNA
ON
FLYING BOAT TYPE H-16-A

FIGURE 33

The skid fin type of antenna is also used on the H-16-A flying boat, consisting of a rectangle, and with a lead-in wire similar to that used on the HS-2-L boat. The antenna constants of this skid fin antenna are illustrated in Figure 34.

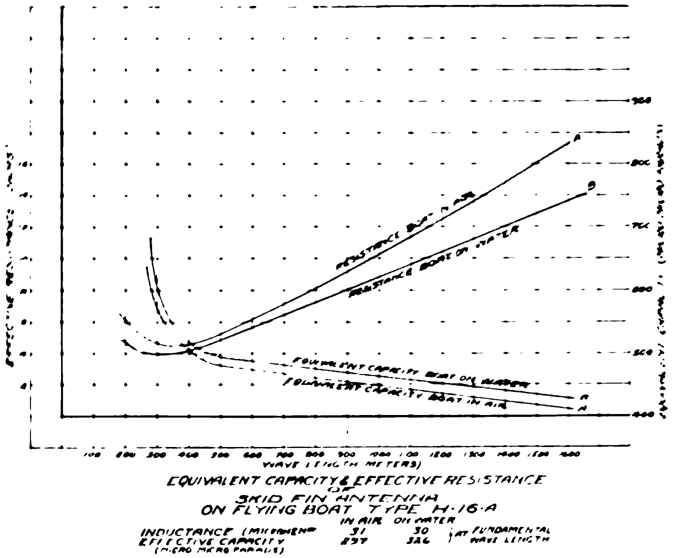


FIGURE 34

Another large type of flying boat is the F-5-L type, illustrated in Figure 35. This boat is approximately the same size as the H-16-A, and differs only in minor structural details. The radio problems connected with the two boats and antenna characteristics may be considered identical. The type F-5-L is a later development and a more standardized form of boat.

The largest flying boat in the naval service is that known as

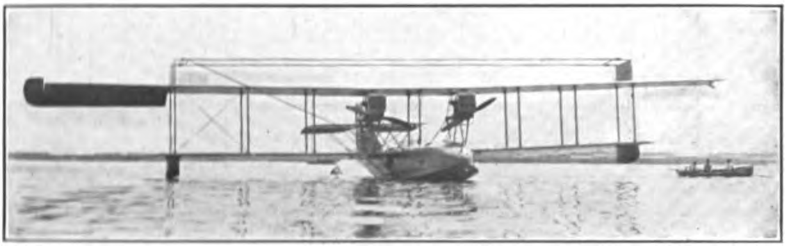


FIGURE 35

the NC type, illustrated in Figure 36. It is this model which was used on the trans-Atlantic flight. Only two types of antenna have been used on this craft, the single trailing wire and the skid fin antenna. On the NC boat the radio station is located in the extreme after-part, and the single trailing antenna passes from the tail almost amidships. The skid fin antenna is arranged similarly to those on the other flying boats, but is somewhat larger and the lead-in is carried aft instead of forward. The antenna characteristics of this boat are not available at the present time.

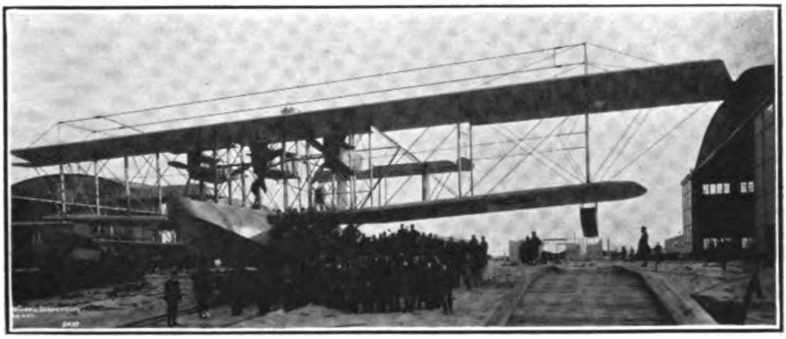


FIGURE 36

Several types of non-rigid dirigibles are used in the naval service as indicated in the table in Figure 12. The dirigible most in use is class "C." illustrated in Figure 37. At first a

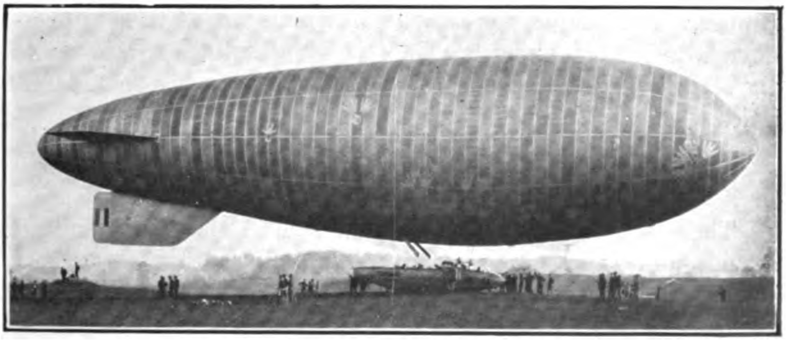
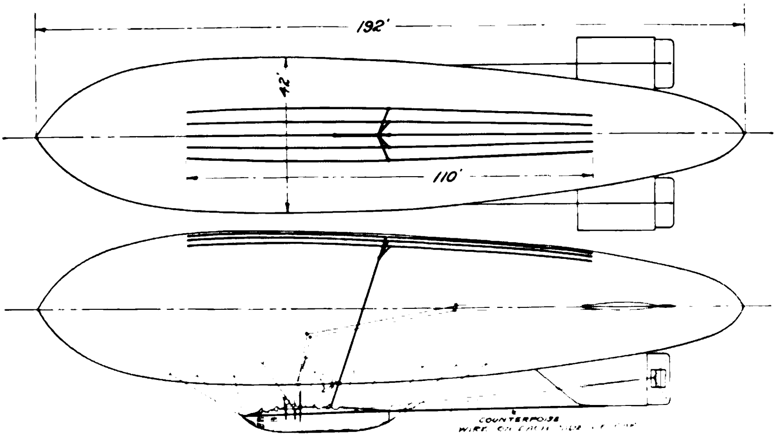


FIGURE 37

single wire trailing antenna was used on all dirigibles but recent investigations have proven the advisability of using a 5-wire "T" antenna suspended within the envelope. The standard form of this antenna, as used in the class "C" dirigible, is illustrated in Figure 38. It is approximately 110 feet (33.5 meters) long, the



ANTENNA INSTALLATION
ON
INTERIOR OF ENVELOPE OF NON RIGID DIRIGIBLE CLASS "C"

NOTE:- WIRES SPACED 3 FEET 6 INCHES APART
AND 11 INCHES FROM TOP OF ENVELOPE.

FIGURE 38

wires being placed about 3 feet 6 inches (1.07 meters) apart and placed, not in one plane, but at a fixed distance, 11 inches (27.9 centimeters) from the top of the envelope. The wire used consists of sixteen strands of number 30 B. and S. gauge copper wire* wrapped with one layer of cotton and one layer of para rubber. The counterpoise used consists of two of these rubber-covered wires run along the sides of the dirigible car and aft to the stabilizer surfaces as illustrated in Figure 38. The characteristics of this radiating system are illustrated in Figure 39.

STANDARD TRANSMITTING EQUIPMENT—SPARK TYPE

There are at the present time in use in the naval aircraft service four types of standard spark radio transmitters. The low-power short-wave spark transmitting equipment for small flying boats, such as type HS-2-L, is known as type SE 1300, the

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* Diameter of number 30 wire = 0.010 inch = 0.25 mm.

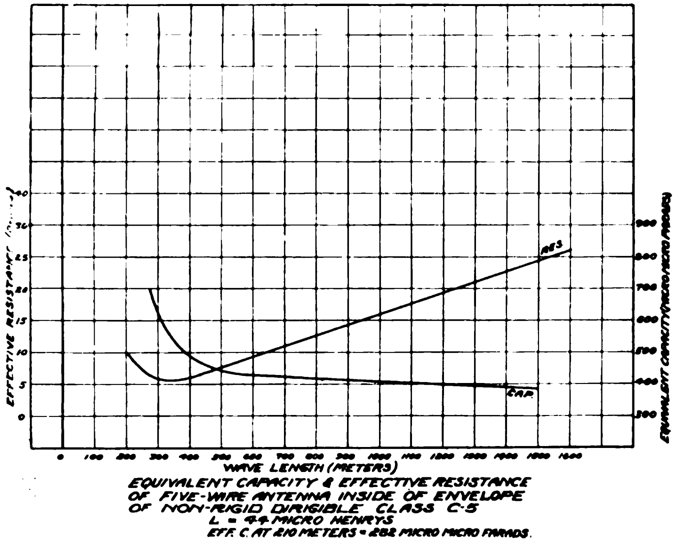

FIGURE 39

main elements and connections of which are illustrated in Figure 40. This set is manufactured by the International Radio Telegraph Company. The equipment completely installed weighs 65 pounds (29.4 kilograms), and will render normal operation from aircraft to shore over all distances within 100 nautical miles (185.9 kilometers). Its important features are its simplicity of control, freedom from operating difficulties, small space

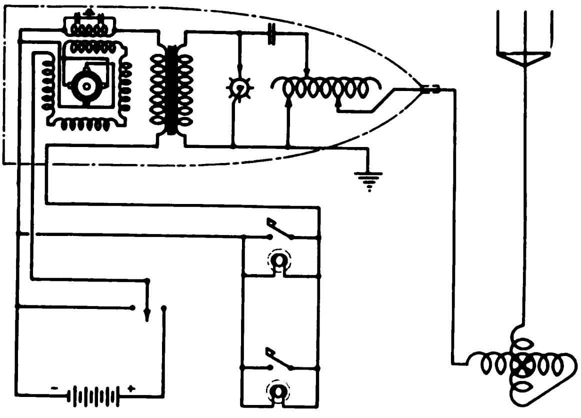

FIGURE 40

occupied, and lightness of weight. The set is of the synchronous rotary spark type consisting of a radio assembly containing the main elements of a spark transmitter, a variometer mounted within the flying boat body for tuning the antenna circuit, send-receive switch, a generator field priming battery, and necessary connections and fittings.

The radio assembly, illustrated in Figure 41, is mounted in a streamline case on the wing of the flying boat, the generator and spark gap, which are mounted in the forward end, being driven by a small propeller. Behind the generator and spark gap are mounted the power transformer and condenser, and oscillation transformer.

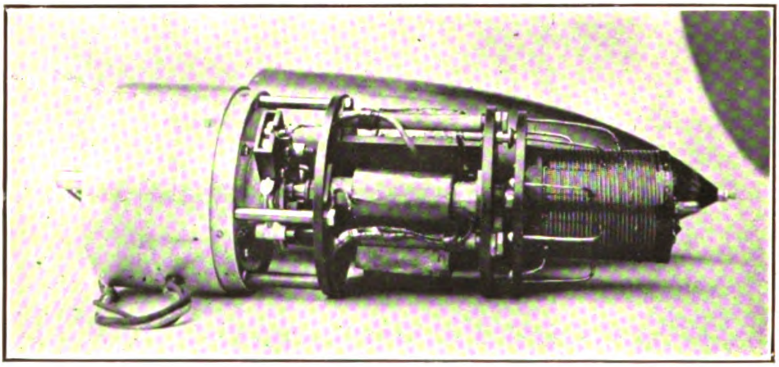

FIGURE 41

The generator is a special inductor type alternator rated at 5,000 r.p.m., 127 to 140 volts on open circuit. The frequency of the generator at normal speed is 1,000 cycles per second. The power transformer is of the closed core type. The mica condenser has a capacity of 0.004 microfarads.

The spark gap consists of a rotary brass disc forming one electrode of the gap and a tungsten electrode forming the other or fixed member.

The oscillation transformer consists of solid bare copper wire wound in grooves around a hollow redmonol cylinder fastened to a redmonol disk which is supported on a short shaft held in the insulating frame of the set. Sixteen and one-half turns of this coil serve as the primary and the remaining eight turns as the secondary, the two coils being inductively coupled with their common point grounded. Nine primary taps are brought out to contact buttons on the disk at the front end of the coil and each button is marked with the corresponding wave length of the closed oscillatory circuit. The available wave lengths are 200, 226, 257, 291, 331, 377, 426, 485, and 550 meters. The stream-line hood which houses the radio assembly is made of canvas bakelite.

The output current from the oscillation transformer is delivered to the antenna circuit thru a special coil spring and socket connector in the tip of the hood which bears on the contactor of the oscillation transformer secondary switch. The antenna current is lead thence from the radio assembly to the body of the boat and thru the variometer and lead-off fittings to the trailing antenna.

The variometer, shown in Figure 42, consists of a cylindrical coil of 87 turns of number 18 B. and S. gauge bare solid copper wire* wound on a thin hollow cylinder of insulating material.

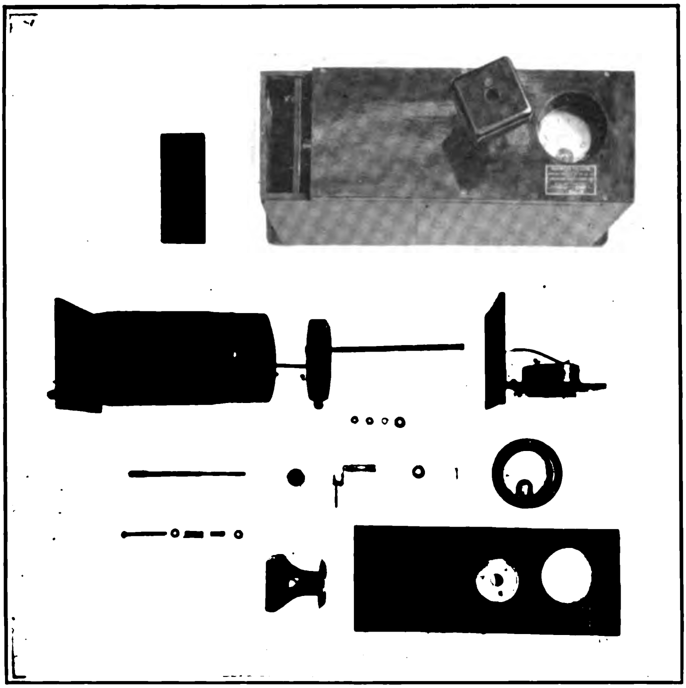

FIGURE 42

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* Diameter of number 18 wire = 0.040 inch = 1.014 mm.

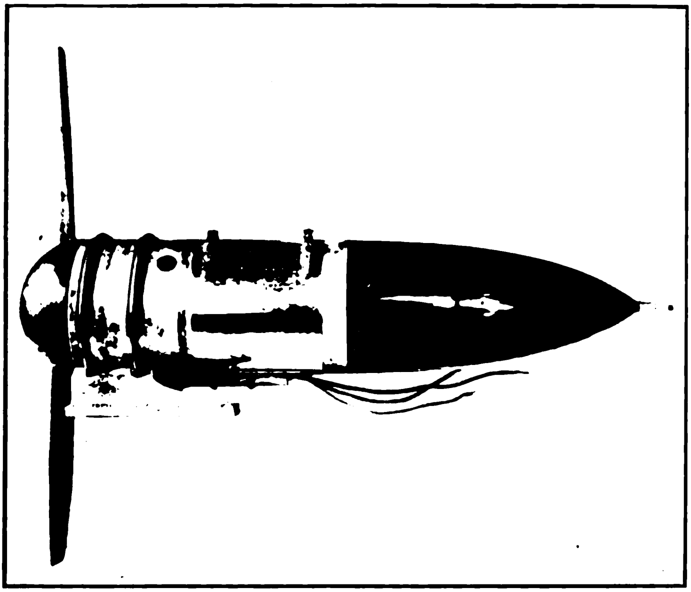
The inductance is varied by means of a trolley mounted within the coil which moves parallel to the axis of the coil, and makes contact with the bare turns of the wire as it passes. It can thus be made to cut out or cut in one turn of the wire at a time as it is moved along. This trolley is moved by means of a handle on the cover of the box which rotates a pinion engaging with a ratchet on which the sliding trolley is fastened. The dead end of the coil is connected to the trolley contact and is thereby short-circuited. A short copper cylinder disc, moving with the trolley inside the coil, serves to prevent sparking at the trolley contact by shielding the magnetic lines of force from the turn short-circuited by the trolley. It serves also to make the changes in the inductance more gradual. The eddy current losses in this disc are negligible. The inductance may be varied between the limits of 35 and 400 microhenrys, and is insulated for 25,000 volts.

The aircraft flameproof keys are used in connection with this set, being connected in parallel, one key in the observer's cockpit forward and the other near the pilot's position. In order to allow either operator to recognize when the other is transmitting, the keys are provided with small incandescent lamps of standard 6-volt type placed in the low tension winding of a small transformer, the high tension winding of which is connected directly across the key contacts. Hence both lamps glow when neither key is depressed and the generator is functioning properly. The transformer lamp and key are all mounted on a single base in a light compact unit.

The send-receive switch is arranged so that as the lever is being thrown to the transmitting position, the generator field is momentarily excited from the priming battery, insuring the building up of the field which is closed without external excitation when the switch is in full transmitting position.

The standard high power short wave transmitting equipment, similar in type to that just described, is known as type SE 1310, also manufactured by the International Radio Telegraph Company. This is also designed for flying boats such as type HS-2-L, and consists of the same essential elements, connected in the same manner, as in Figure 40. This equipment, completely installed, weights 77 pounds (34.9 kilograms), and will render normal operation from aircraft to a shore station over all distances within 300 nautical miles (598 kilometers). This range is specified for normal operating conditions only. Far greater range may easily be obtained with this equipment as demonstrated by the trans-Atlantic flight results, described later.

The radio assembly, illustrated in Figure 43, is of the same general design as that of the 200-watt set, but is somewhat larger and contains several slight differences in constructional detail.



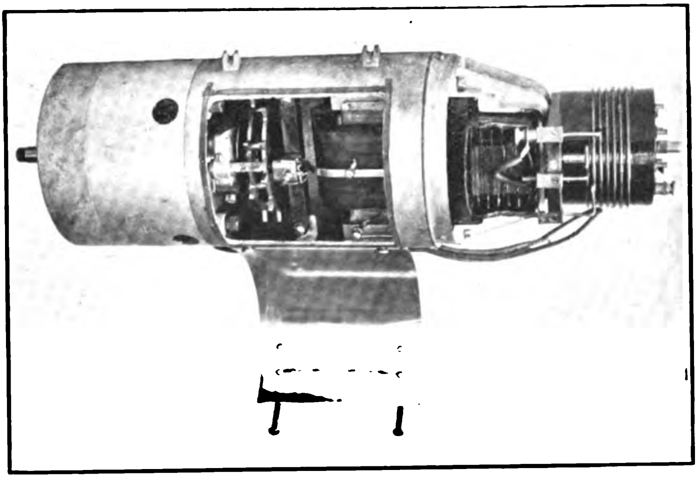


FIGURE 43

The generator is a special inductor type alternator rated at 5,000 r.p.m., no-load voltage 135-150, full-load voltage 90, 500 watts. The frequency of the generator at normal speed is 1,000 cycles per second. The power transformer is of the closed core type. The mica condenser has a capacity of 0.010 microfarads.

The spark gap consists of a combination eight-tooth and twelve-tooth rotor mounted on an insulating hub, with a tungsten electrode arranged for operation with either rotor. This spark gap arrangement enables the operator to adjust the spark frequency for 1,000 sparks per second or 666 sparks per second, without changing the gap rotor.

The oscillation transformer consists of solid bare copper wire wound in grooves around an insulating cylinder fastened to the condenser frame. The oscillation transformer is of the usual conductive coupled type having the coupling turns common to the primary and antenna circuits, and having the common point grounded. There are seven turns total, with taps brought out to give 335, 375, and 425 meters respectively. Coupling is provided by bringing out taps from the three turns adjacent to the ground end of the coil. These taps terminate in a suitable switch provided at the end of the oscillation transformer to permit proper coupling adjustments.

The output current from the oscillation transformer is delivered to the antenna circuit in the same manner as in the case of the 200-watt set.

The variometer is of the same general construction as is indicated in Figure 42, consisting of a cylindrical coil of 50 turns of number 14 B. and S. gauge bare copper wire*. The inductance may be varied from 120 microhenrys to approximately zero.

Two aircraft flameproof keys with indicator lamps and a standard send-receive switch are used, as in the case of the 200-watt set.

Another type of standard high power short wave transmitting equipment supplied for installation on flying boats is known as type SE 1320, and is designed and manufactured by Cutting and Washington, Incorporated. This equipment, completely installed, weighs 77 pounds (34.9 kilograms), and will render normal operation from aircraft to shore over all distances within 200 nautical miles (370 kilometers). The main elements and method of connection are illustrated in Figure 44. The trans-

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* Diameter of number 14 wire = 0.064 inch = 1.625 mm.

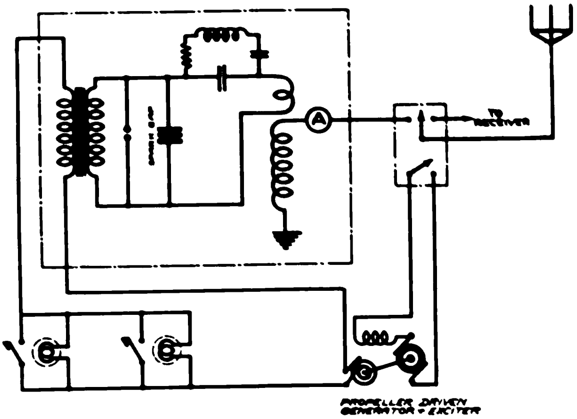

FIGURE 44

mitter is of the impact excitation type[1] and is designed for transmission on a wave length of 375 meters only.

The main element of the set, a transmitter panel, 12 inches (30.5 centimeters) wide, 17 inches (43.2 centimeters) high, 10.5 inches (26.6 centimeters) deep over-all, and weighing 15 pounds (68 kilograms), is illustrated in Figure 45. This is made up of

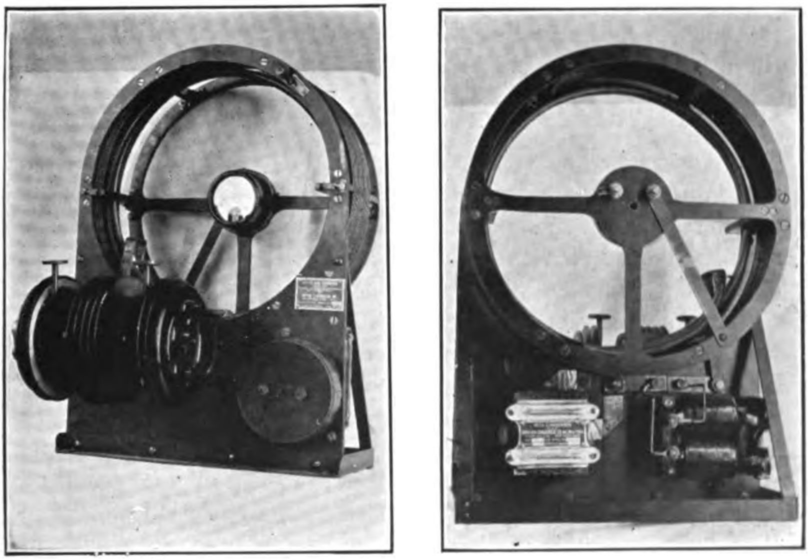

FIGURE 45

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[1] See Bowden Washington, "On the Electrical Operation and Mechanical Design of an Impulse Excitation Multi-spark Group Radio Transmitter," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, 1918, volume 6, page 295.

a small formica panel with a formica ring at the back supporting the primary and secondary coils, ammeter, transformer, spark gap, condenser, and concentration circuit, the whole being assembled on a mounting base arranged for fastening down to a shelf in the body of the boat. On the front of the panel below the oscillation transformer are the spark gap and concentration circuit inductance. At the rear below the oscillation transformer are the primary condenser, safety gap, concentration circuit, and transformer. On the rear helix posts supporting ring is the antenna ammeter reading from zero to 5 amperes.

Power is supplied from a propeller-driven alternating current generator contained in a stream-line case and mounted on one of the wings of the craft. The generator exciter and armature are mounted on the same shaft. The exciter is bi-polar and delivers 110-volt direct current to the generator field. The generator is of induction type with eight projections on the rotor and four stator coils in the eight slots delivering 220 volts, 500 cycles at no load. This generator was designed and manufactured by the Crocker-Wheeler Company.

The spark gap consists of two tungsten gaps in series, separately adjustable and with heat radiating fins for cooling. A locking device is provided so that after adjustment a gap can be locked in operating position. Flameproof keys with indicator lamps and a send-receive switch are used as in the case of the other spark transmitters.

STANDARD TRANSMITTING EQUIPMENT—VACUUM TUBE TYPE

In describing the vacuum tube radio transmitters used on naval aircraft, it is of interest to first consider the standard types of tubes utilized. Among the four standard transmitting tubes used it will be noted that one has a coated cathode filament and the others tungsten filaments.

The question of a "Wehnelt" or coated cathode filament versus tungsten filament and electrodes for vacuum tubes is similar to many engineering questions, which involve the development of two different devices or two different methods to fill substantially the same need. Each device or method has its advantages as well as its disadvantages, the best device or method to use depending upon the many conditions of service. As in almost all questions of this kind, so in this case, each type of filament and construction will find and fill its own field. At the present time these features may be classified as follows:

The coated filament type of tube has these advantages:

(1) For the same filament current and voltage, the filament is of greater length (3 to 6 times), this increased cathode area facilitating the design of a tube of low impedance. Other factors remaining constant, the lower the tube impedance, the greater the energy amplification possible. This is of particular value when the tube is used as an amplifier or modulator.

(2) For a given life and total electron emission the coated filament requires less energy for filament excitation. This energy ratio in the small transmitting tubes described (Type CG 1162) would be about 1 to 1.5.

(3) Low initial electron velocity. This aids in the design of receiving tubes to have a very high input impedance. This factor is of importance in a tube used for the amplification of small voltages, that is, for amplifying faint received signals.

The coated filament has the following disadvantages, most of which are due to the fact that the filament is not of a homogenous material:

(1) The coating of applied material may tend to become irregular, causing so-called "hot spots," making the life of such tubes short.

(2) The so-called secondary emission or "blocking," owing to the nature of the emitting material, is on occasions a most mysterious and baffling difficulty preventing the functioning of the tube.

(3) The lower filament temperatures, together with the non-homogeneity of the filament, increases the likelihood of contamination of the filament surface, making its manufacture and use more critical to degree of vacuum and lowering the uniformity of the product.

The tungsten type of tube has these advantages:

(1) Ease of both obtaining and maintaining a high vacuum.

(2) All parts may be operated at a comparatively high temperature continuously which, with the use of a very high voltage direct current, makes practical the production of high power tubes of reasonable design operating at very good efficiencies.

(3) Owing to the comparative ease of handling and exhaust treatment, the quantity production of a com-

mercial number of tubes is a better manufacturing proposition. This factor in favor of the tungsten tube increases with the power of the tube.

(4) Tungsten filaments are made of a very pure metal and tungsten is a very stable material, so that the uniformity of the manufactured tube is very high, mechanical dimensions and spacing being the only factors.

The tungsten tube has the following disadvantages:

(1) The cathode having smaller area necessitates close mechanical work to secure a tube of low impedance.

(2) The variation of electron emission with temperature is greater, therefore the output of a tungsten power tube varies more with change of filament current. In certain cases this feature of less inherent regulation is an advantage as it allows a tube to be forced for high output if a very short life can be tolerated. It is also a property that is taken advantage of in the construction of "kenotrons" for use as regulator tubes.

It is believed that a study of these features will show that for receiving tubes and very low power transmitting tubes the coated filament type may have a balance in its favor. For transmitting tubes, from five watts up, the balance is probably in favor of the tungsten tube and its advantages relatively become greater, the higher the power of the tube.

The essential data on the various transmitting tubes utilized is tabulated in Figure 46.

Tube type CW 931, illustrated in Figure 47, is a coated filament tube, manufactured by the Western Electric Company. Its characteristic is shown in Figure 48.

Tube Type CG 1162, illustrated in Figure 49, is a tungsten filament tube, manufactured by the General Electric Company. Its characteristic is illustrated in Figure 50.

Tube type CG 1144, illustrated in Figure 51, is a tungsten filament tube, manufactured by the General Electric Company. Its characteristic is illustrated in Figure 52.

Tube type 916, illustrated in Figure 53, is a tungsten filament tube, manufactured by the General Electric Company. Its characteristic is illustrated in Figure 54.

Navy Type Number	CW 931	CG 1162	CG 1144	CG 916
Signal Corps Type Number	VT-2	VT-14	VT-18	None
Manufacturer	Western Electric Company	General Electric Company		
Manufacturer's Type Number	E	T	U	P
Watts Output	5	5	50	250
Filament Volts	7.0	7.5	10.0	18.0
Filament Amperes	1.35	1.75	6.5	3.6
Plate Volts	350	350	750–1000	1500
Plate Amperes	0.040	0.040	0.150	0.250
Length over-all	4.25 inches 10.8 cm.	4.31 inches 10.92 cm.	7.5 inches 19.0 cm.	14.25 inches 39.15 cm.
Diameter, Maximum	2.28 inches 5.78 cm.	1.75 inches 4.44 cm.	2.0 inches 5.08 cm.	5.0 inches 12.7 cm.
Weight	1.87 ounces 53 grams	1.86 ounces 52.6 grams	5.88 ounces 166.5 grams	23.31 ounces 660 grams

FIGURE 46—Transmitting Tubes

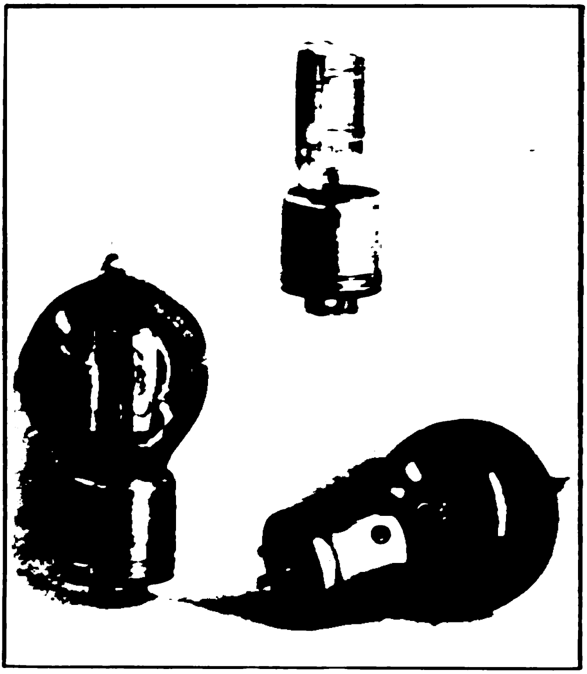


FIGURE 47

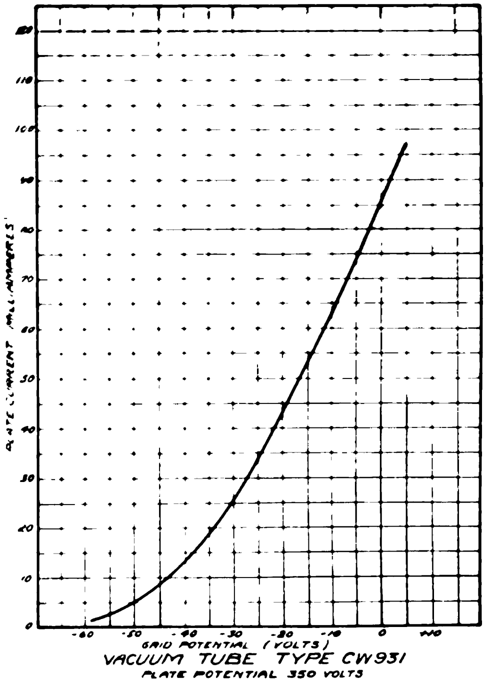


FIGURE 48

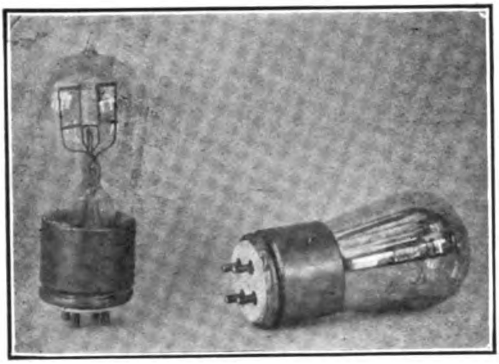


FIGURE 49

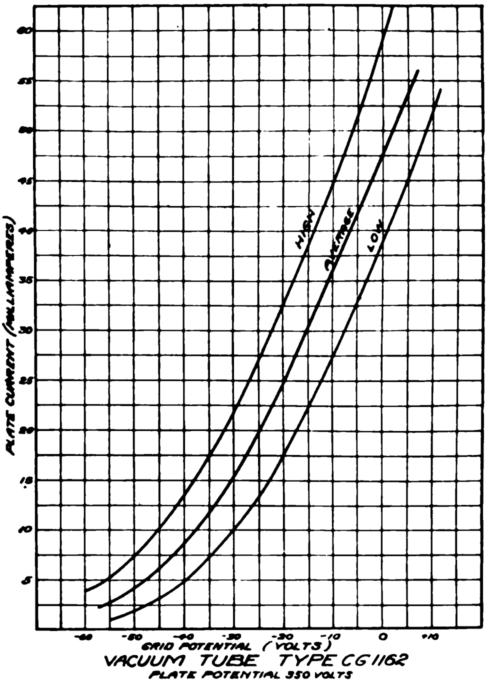


FIGURE 50

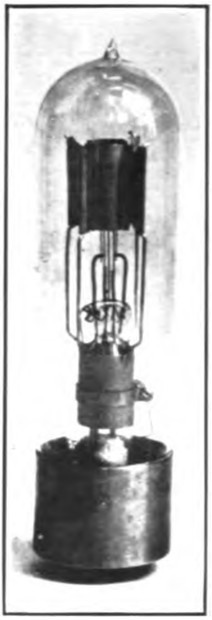


FIGURE 51

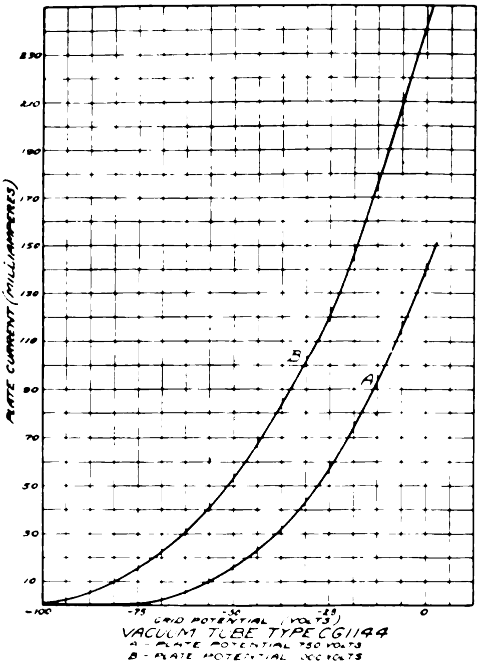


FIGURE 52

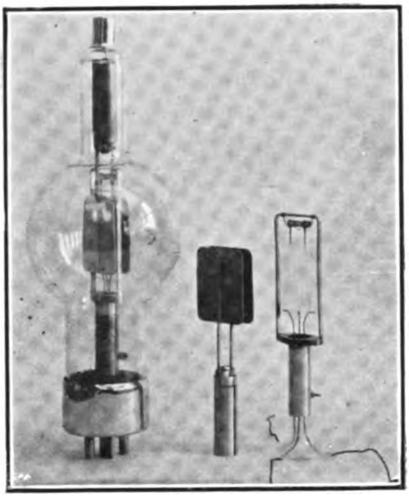


Figure 53

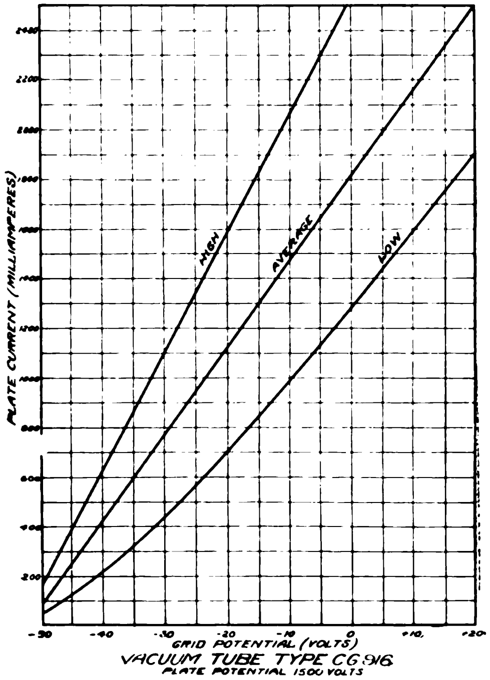


Figure 54

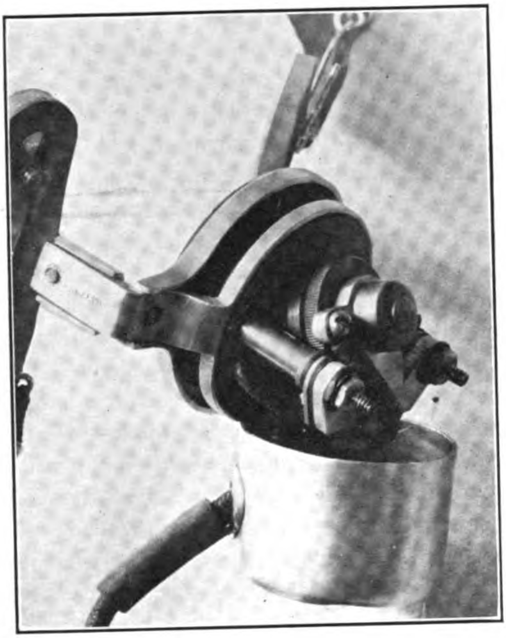
In the tabulated data on the above tubes it will be noted that no numerical figures of life are given. This is because, as in an incandescent lamp, life and efficiency are opposing factors, and life is a matter of engineering judgment rather than a property. For aircraft work where weight efficiency becomes very important, a short life suffices if a tube is operated at a very high efficiency.

In the production of radio telephone sets for naval aircraft it was found necessary to develop a microphone transmitter which would operate satisfactorily under the terrific noise conditions existing on naval seaplanes and flying boats. Any standard microphone transmitter was absolutely useless, and even those which had already been developed for aircraft work did not operate with a high enough degree of satisfaction, especially for long distance transmission.

After a period of experimenting on the part of several manufacturers, the Magnavox Company of San Francisco, California, solved the problem most satisfactorily by a transmitter of the design illustrated in Figure 55. In this construction the diafram is so mounted that it is exposed on both sides to extraneous noise vibrations, and thus is only actively affected by the directional impulses of the voice from one side of the diafram. In order to prevent the carbon granule from falling away from the front electrode when the transmitter is held in a horizontal position, as when the observer is looking over the side of the boat, the button is mounted at an angle of 30 degrees with the diafram, which insures operation in nearly any position in which the pilot or operator is apt to place himself.

The transmitter is supported in front of the mouth by two light silk-covered rubber straps fastened to the helmet. As the transmitter is very light, weighing only three ounces (85 grams), this means of support causes no inconvenience, and is of very decided advantage in that the transmitter is always in position for use. All connections, as well as the carbon granule button, are enclosed in a light aluminum spray-proof case.

In the naval aircraft service, radio transmission from the craft while resting on the water is of great importance, especially in cases where the aircraft has become disabled and assistance must be summoned. At such a time there would be no power available from a propeller-driven generator. In the design, therefore, of tube transmitters, a new arrangement for power supply was made. This consisted of a storage battery, floating across the leads of a direct current generator driven directly



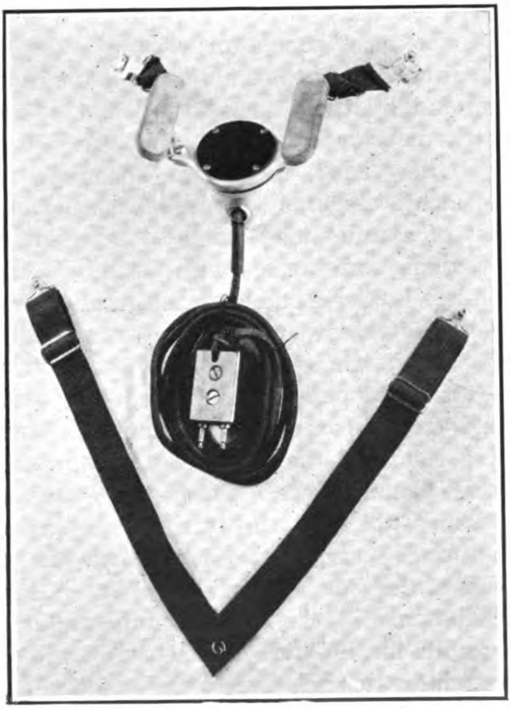


FIGURE 55

by the aircraft engine. In this way whenever the aircraft is in flight the radio power is derived as directly as possible from the primary source of power on the craft, and the engine-driven generator will assist in keeping the storage battery charged for use at any time when the engines are not running. With a standard engine generator connected to each engine, the higher power radio sets on the larger craft are properly supplied from additional generators available due to the additional engines used.

In view of the above arrangement the standard tube sets described herein are given a time rating of operation based on alternate five-minute periods of transmission, and ten-minute periods of reception, power being taken from a standard storage battery consisting of 63-pound (28.6-kilogram) 12-volt units. In the case of the 12-volt equipments a single unit is assumed, and in the case of 24-volt equipments two units connected in series are assumed unless otherwise specified. An arbitrary rating of one hour's operation therefore may be interpreted as four five-minute periods of transmission with four intervening ten-minute periods of reception with a standard receiver. All of the 12-volt storage battery units supplied for naval aircraft radio are of the lead plate type.

Several of the most successful vacuum tube transmitters used on naval aircraft, now to be described, have been developed and manufactured by the General Electric Company. While the transmitters developed by that company differ greatly in many respects, they have, in general, the same fundamental circuit arrangement. Figure 56 shows schematically this fundamental circuit of a radio telephone and telegraph transmitter capable of providing radio communication by either continuous wave telegraphy, buzzer-modulated telegraphy, or telephony.

The elements of the vacuum tube generating the high frequency energy are designated by "F," "G," and "P" representing filament, grid, and plate respectively. The grid and plate circuits are inductively coupled to the antenna circuit by means of the antenna coil "L," the grid coil "L_2" and the plate coil "L_1." With the filament lighted and plate circuit closed, feeble natural oscillations will be set up in the antenna circuit due to the slight transfer of energy between the plate and antenna systems. Because of the inductive relation of the grid and antenna coils these oscillations will be impressed on the grid of the vacuum tube. This results in amplified oscillations in the plate circuit which, thru its inductive relation to the antenna system, reinforces the oscillating energy in the antenna circuit. This action is repeated with the antenna current constantly increasing until it is limited by the antenna and tube characteristics. The values of the grid leak "*R*" and grid leak condenser "C_2" are determined principally by the tube, and are designed to permit the tube to operate on the proper portion of its characteristic curve. The design of the coil system depends upon the wave length, antenna, and tube characteristics. The condenser "C_1" acts as a radio by-pass and also as a protective condenser for the generator. The condenser "*C*" functions as a radio by-pass for the oscillating current in the plate circuit. The modulation indicator "*M*" and the reactance "L_3" will be considered later.

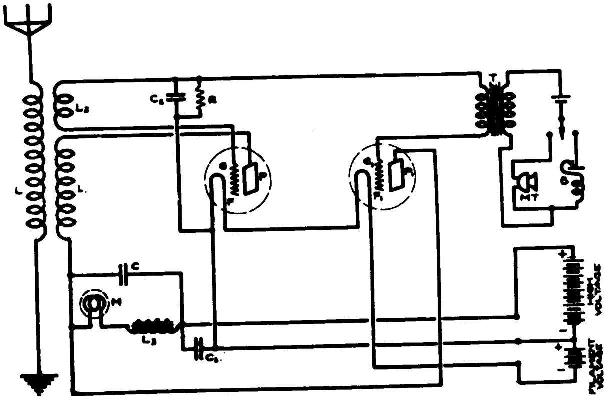


FIGURE 56

For use in continuous wave telegraphy the telegraph key is inserted in the circuit so as to open the grid leak or else to remove a short circuit from a second condenser in the grid circuit. The effect is the same in both cases, namely to stop the tube from oscillating. In aircraft radio, the first method is usually desirable in that an additional condenser is not necessary, hence decreasing the weight.

The elements of the vacuum tube employed as a modulator are represented by "F_1," "G_1," and "P_1." The microphone transformer is shown as "*T*." The buzzer "*B*" and the microphone transmitter "*MT*" are interchangeable by means of suitable switching. The modulation of the radio frequency energy generated by the oscillator tube is accomplished by means of amplifying the output of a microphone transformer, the primary of which is in series with a microphone and a source of direct current. The secondary of this transformer is connected to the grid of a vacuum tube termed the modulator, the grid being maintained at the proper operating point on its characteristic by a sufficient amount of negative potential. The output of the modulator is introduced into the plate circuit of the oscillator thru the incandescent lamp "*M*," or modulation indicator, and the iron core reactance "L_2." The brilliancy of the lamp "*M*" is varied by the fluctuations in plate current, hence indicating the functioning of the modulator tube. When used as a radio telephone transmitter suitable switching connects in the microphone and short circuits the telegraph key. When used for modulated buzzer telegraphy a buzzer replaces the microphone and the telegraph key controls the oscillations of the tube oscillator as in continuous wave telegraphy.

The lowest power vacuum tube set, developed by the General Electric Company, is a combined transmitting and receiving equipment known as type SE 1345. This equipment is used on a small ship airplane for fire control work in connection with battleships, the airplane being of the one seat type so that the radio set must be operated by the pilot. The important features are smallness, lightness of weight, and simplicity of control, it being merely necessary to operate the send-receive switch after initial adjustments are made. The apparatus gives an antenna input of 5 watts and is designed for telephone communication with a battleship within an operating radius of thirty nautical miles (55.5 kilometers). The main element of this equipment, as shown in Figure 57, is a cabinet 14.38 inches (36.4 centimeters) wide by 5.38 inches (13.6 centimeters) high by 7.5 inches (19.0 centimeters) deep, weighing 12 pounds (5.44 kilograms).

The transmitter employs two vacuum tubes, type CG 1162 one tube being used as an oscillator and the other as a modulator. The receiver employs three receiving tubes, type CW 933, one being used as a detector, and the others in connection with two stages of audio frequency amplification. The tubes are mounted on a rubber cord suspension. It has been found that while rubber cord suspension is more satisfactory for receiving apparatus or combined sets of low power, spring supports are preferable for transmitting tubes. The plate circuit power and filament current for the transmitting tubes are supplied from a propeller-driven generator, illustrated in Figure 58. Constant rotary speed for this generator thruout varying airplane speeds is accomplished by means of a self-regulating propeller. The filament current for the receiving tubes is supplied from a 4-volt storage battery which will permit of reception for a period of four hours.

The schematic diagram of connections is illustrated in Figure 59. This equipment is arranged for operation on one wave length only, merely sufficient inductance variation being pro-

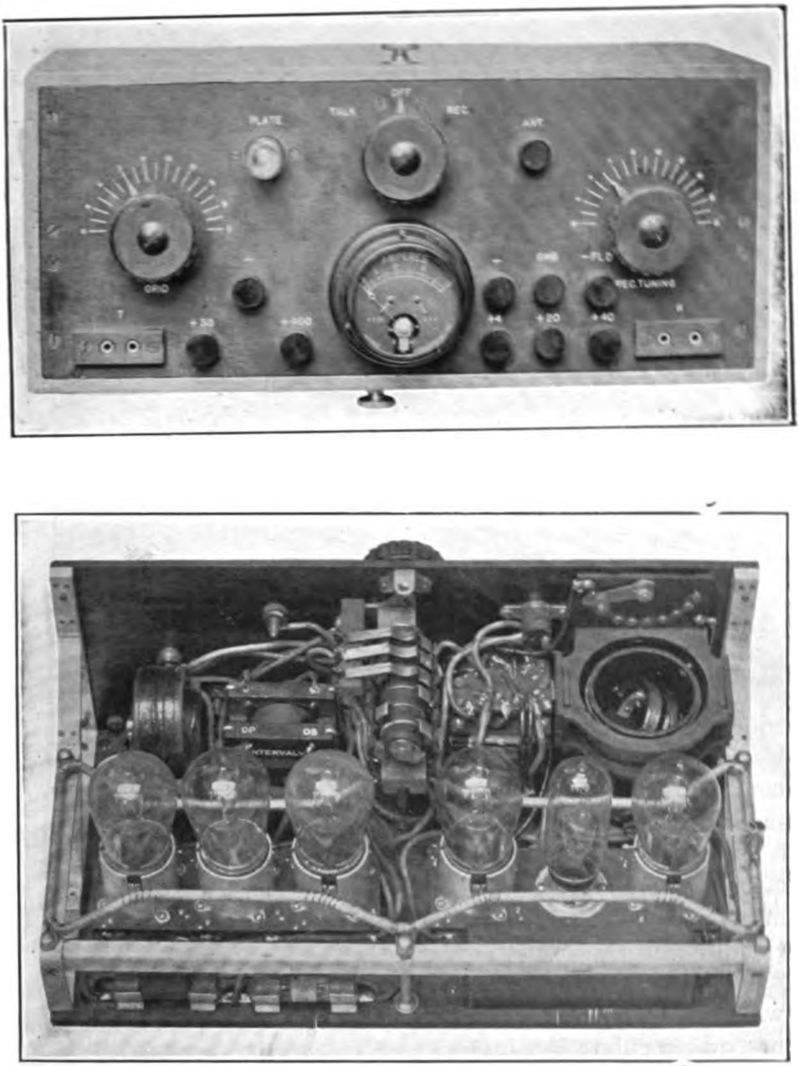


FIGURE 57

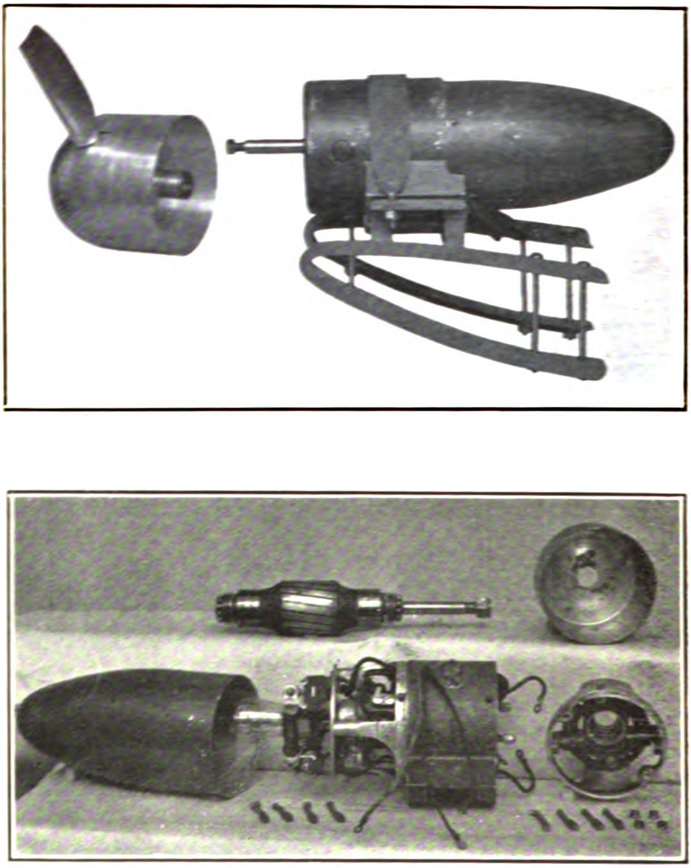


FIGURE 58

vided for accurate adjustment to the antenna utilized. The wave lengths used vary between 450 and 600 meters. The controls, mounted on the front of the cabinet panel, consist merely of a send-receive switch, a transmitter grid coupling adjustment, and a receiver variometer for tuning. On the front of the panel is also mounted an antenna ammeter and a small incandescent lamp in the transmitter plate circuit as a modulation indicator.

The microphone transformer is provided with a side tone winding so that the pilot may hear his own speech and thus properly regulate the intensity of his voice. An iron filament ballast lamp, placed in series with the filaments of the two transmitting tubes, is used to maintain constant filament current of 1.75 amperes throut all operating ranges of generator speed. An important feature of this equipment is that the panel and entire apparatus mounted behind it may be removed from the containing case without disconnecting the leads and rendering the set inoperative.

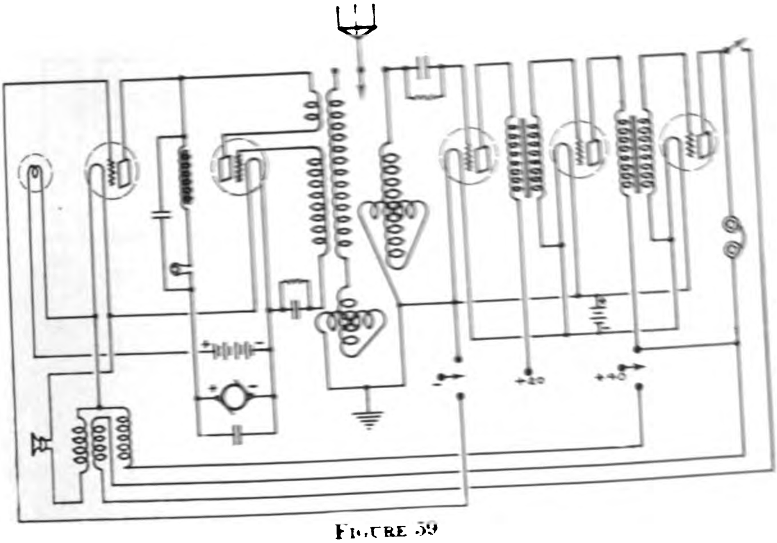


FIGURE 59

This equipment is arranged for use on an antenna composed of the two sides of the metal portions and brace wires of the plane, or in connection with a light trailing antenna used against the entire metal portion of the plane as a counterpoise. In the latter case, a special release is provided by which the trailing wire may be cast off with but a slight movement on the part of the pilot.

Another of the low power vacuum tube transmitters used is that embodied in combined transmitting and receiving equipment, type CW 1058, an adaptation of Signal Corps type SCR 68 equipment to naval uses. This apparatus is manufactured by the Western Electric Company. In this case the power for the transmitter plate circuits is supplied from a dynamotor driven from a 12-volt storage battery instead of from a propeller-driven generator as in the case of the Signal Corps equipment. The transmitter employs two low-power transmitting tubes, type CW 931, one used as an oscillator and the other as a modulator. The receiver consists of a detector and two stages of audio frequency amplification, three receiving tubes, type CW 933, being used. A schematic circuit diagram of this equipment is illustrated in Figure 60. The cabinet, illustrated in Figure 61, is approximately 17 inches (43.1 cm.) wide, 10 inches (25.4 cm.) high, 7 inches (17.8 cm.) deep, and weighs 21 pounds (9.5 kilograms).

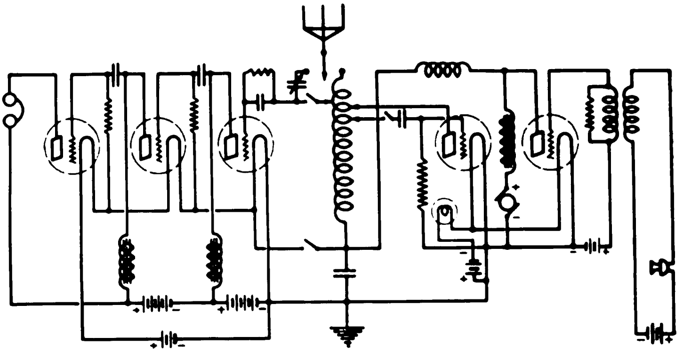


FIGURE 60

The dynamotor requires a current of 12 amperes from the 12-volt battery, the transmitting tubes a filament current of 2.7 amperes, and the receiving tubes a current of 3.3 amperes. A total current of 14.7 amperes is thus drawn from the battery while transmitting, and a current of 3.3. amperes while receiving. A standard 63-pound (28.6-kilogram) battery, fully charged, will permit continuous operation for a period of three hours on a basis of alternate five-minute periods of transmission and ten-minute periods of reception. The dynamotor is mounted on a case holding the receiving plate batteries, the entire dynamotor unit weighing 19 pounds (8.6 kilograms).

This equipment is arranged for transmission or reception on wave lengths of 215 to 450 meters using a trailing antenna. The antenna coil is oval shaped and is used for both transmitting and receiving. The transmitting tubes are mounted within this coil, an arrangement originally made to reduce the range of the set for Army uses and to economize space. Taps on this coil are connected to the dial switches marked "Wave length" "Coupling," and "Input" located in the penthouse on the front of the panel. By means of the dial switch marked "Input," the voltage across the [illegible] proper grid voltage to make the [illegible] of the [illegible] the [illegible] as to secure the maximum power [illegible] time the indicators [illegible] "Wave length" to give the [illegible] this equipment [illegible]

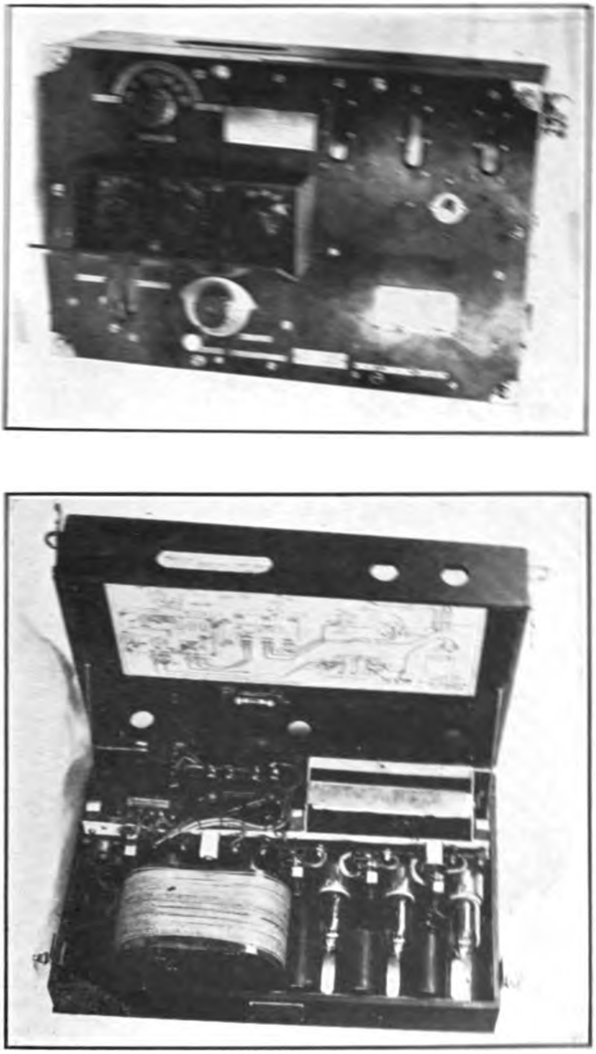


FIGURE 61

can be maintained between aircraft and ground station up to a distance of 10 nautical miles (18.5 kilometers), and between aircraft up to a distance of five nautical miles (9.3 kilometers).

(*To be continued*)

LONG DISTANCE RADIO COMMUNICATION IN CHILE*

BY

E. W. FIELDING

(RADIO TELEGRAPHIC INSTRUCTOR IN THE CHILEAN NAVY, VALPARAISO, CHILE)

The high power stations of Llanquihue and Punta Arenas are practically identical, each station being provided with 100-kilowatt and 5-kilowatt Marconi apparatus. The following brief description will give a general idea of the equipment.

Both stations are fitted with a 220-horse power, three-cylinder, Diesel engine running at 200 revolutions per minute, and coupled to a 250-volt, 600-ampere dynamo. A 170-brake horse power, 220-volt motor runs in conjunction with a 100-kilowatt, 220-volt, 200-cycle alternator running at 1,500 revolutions per minute, and this in turn is coupled to a 16-stud rotary disc set, thus producing a spark frequency of 400 per second. The battery consists of 122 accumulators; capacity approximately 2,000 ampere-hours.

The 100-kilowatt set is arranged to give wave lengths between 2,400 and 5,000 meters, while the 5-kilowatt set ranges between 600 and 1,600 meters. All the antenna insulators are of porcelain, the leading-in ones containing oil. The earth wires are disposed in the form of two semi-circles; the radius for the circle being 200 feet (61 m.).

There are four antennas.

Antenna A consists of 20 wires, 2,000 feet (610 m.) long.
" B " " 12 " 1,475 " (450 m.) "

Receiving antenna has two wires 1,800 feet (549 m.) long, and the 5-kilowatt set has the usual four-wire antenna.

These antennas are arranged between seven sectional steel masts each 253 feet (77 m.) high; A, B, and receiving antennas being in the form of inverted L's, directive to Punta Arenas, and vice versa. Measuring the fundamental wave length of the two-wire receiving antenna by an atmospheric spark direct to earth, during a storm, gave a result of 2,650 meters.

For reception, the Marconi long wave receiver is used, tuning up to 7,000 meters, and employing carborundum and the Fleming valve as detectors.

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* Received by the Editor, February 3, 1919.

Figure 1 shows the masts and buildings of one of the stations, Figure 2 the Diesel engine, Figure 3 the storage battery, and Figure 4 the remote-controlled antenna switch and the receiving set.

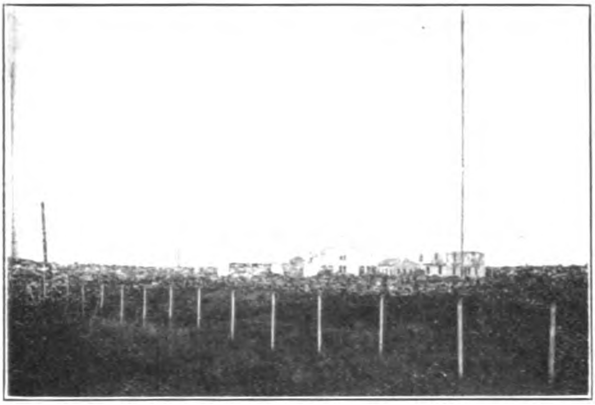

FIGURE 1

The climate at Llanquihue (41° south latitude) is very damp: rain falls very frequently thruout nine months of the year, and during the remaining three months, is generally about to do so. This dampness has caused considerable trouble with the apparatus at various times, generally in the winter, and during this season the buildings have to be continuously heated.

The oil-filled condensers have worked well, only two giving out during three years, the oil having been boiled previously to use.

Punta Arenas and Llanquihue are about 875 miles (1,349 km.) apart. Tests carried out proved that best all-round results, taking into consideration the energy used, were obtained with the wave length of 3,600 meters working with 50 kilowatts. Using 100 kilowatts at 5,000 meters, stronger signals were recorded, but not in proportion to the increase of energy. The 3,600-meter wave, requiring but little added antenna inductance, since it is nearer to the fundamental than the 5,000-meter wave, is now used for all traffic.

Work thruout the winter can be carried on with but small interruption from strays, but during the summer months signals between two and four of the afternoon are considerably weakened, at times dying out altogether. This same effect is also noticed occasionally during the morning, from ten to eleven, the changes in signal strength taking place very quickly. The same phenomena are also noticed in Punta Arenas when receiving from Llanquihue.

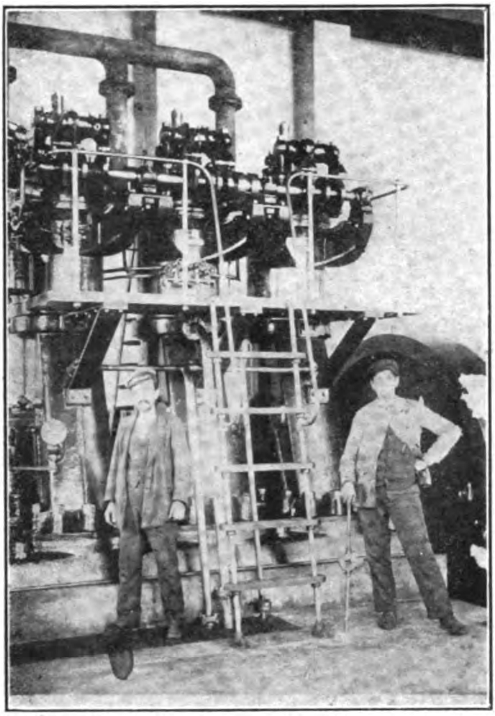

FIGURE 2

From twilight to dawn, signals from Punta Arenas are quite strong when using the carborundum detector, strays as well being at a maximum during this period. Except for hot summer weather or during squalls, these strays can be cut out sufficiently to permit reception. Balanced crystal working gives best results, and by selecting the crystals employed, strays can be heard as one continuous blur; the signals being of different tone can then be easily read. The Fleming valve used in a partially insensitive position also gives good results.

It will be noticed that these two stations lie very nearly on the same degree of longitude. Everything goes to prove that the weakening of signals produced when working from East to West is the same as when working from North to South where the stations enter the daylight zone at nearly the same time.

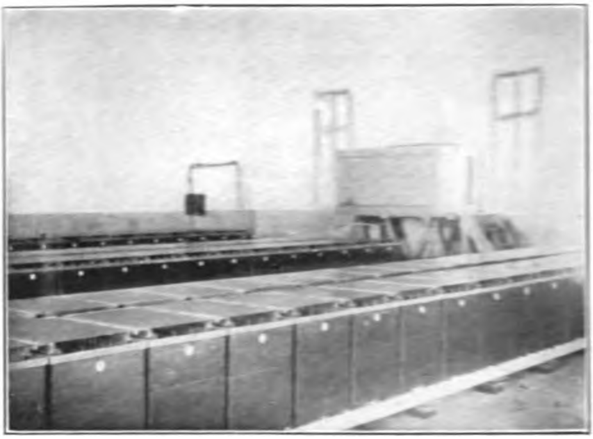

FIGURE 3

During the eruption of "Mount Calbuco," situated about 60 km. (40 miles) from the Llanquihue station, a considerable increase in atmospherics was noticed, and great difficulty in working resulted. At night time, San Francisco and Honolulu can easily be heard but read with difficulty owing to the atmospherics. Signals are improved by using silicon, and employing a very weak current, but extra strong atmospherics will cause it to lose its sensitiveness.

These stations constitute a considerable improvement in communication for this country; and Chile now possesses an uninterrupted chain of radio stations stretching from Arica to Punta Arenas, thus ensuring quick and ready communication.

Needless to say, they have already proved of great utility; and the rapid exchange of radio messages in connection with the Shackleton expedition resulted in the rescue of the seamen stranded on Elephant Island.

With the advent of the vacuum tubes, communication will be made all the more secure and communication with distant countries all the more certain.

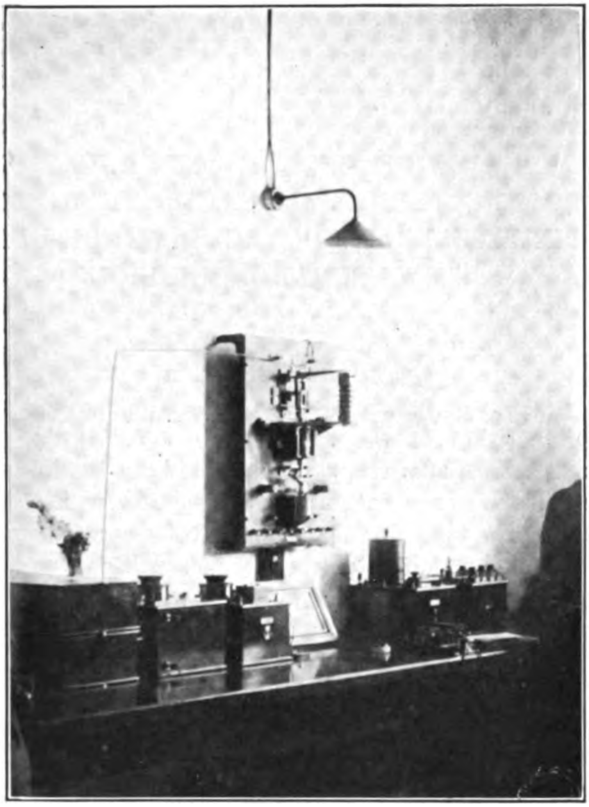

FIGURE 4

SUMMARY: The Chilean radio stations at Llanquihue and Punta Arenas are described in detail. The best wave length and power for this 875-mile (1,349-km.) transmission are given. The effects of strays, their reduction, and signal fading are discussed.

THE DEPENDENCE OF THE AMPLIFICATION CONSTANT AND INTERNAL PLATE CIRCUIT RESISTANCE OF A THREE-ELECTRODE VACUUM TUBE UPON THE STRUCTURAL DIMENSIONS*

By

John M. Miller

(Bureau of Standards, Washington, District of Columbia)

In an earlier paper[1] the writer pointed out the importance of the amplification constant and internal plate circuit resistance in determining the behavior of vacuum tubes as amplifiers and outlined a simple and direct method of measuring these quantities. In the present paper, formulas based upon theoretical considerations are derived which relate the values of these quantities to the dimensions and spacing of the three electrodes.

The two-electrode tube has been treated theoretically by Child[2] and Langmuir.[3] The latter author has derived expressions for the electron current flowing between the filament and plate of such a tube as limited by the space charge in two particular cases and has shown the general law of dependence of the current upon the voltage between the electrodes. In one of the particular cases the filament was assumed to be an infinite plane emitting electrons and the plate likewise an infinite plane parallel to the filament at a distance x and maintained at a positive potential with respect to the filament. In the other case the filament was assumed to be a cylindrical wire of infinite length surrounded by a concentric cylindrical surface of infinite length constituting the plate.

The limitation of the current by the space charge comes about in the following manner. Those electrons which are in motion from the filament to the plate constitute a space distribution of negative electricity which exerts an electrical force in the region of the filament opposite to that due to the plate. No

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* Received by the Editor, April 26, 1919.

[1] Proceedings of The Institute of Radio Engineers, volume 6, page 141, 1918.

[2] C. D. Child, "Physical Review," 32, page 498, 1911.

[3] I. Langmuir, "Physical Review," 2nd series, 2, page 450, 1913.

matter how many electrons are emitted by the filament per second, the current flow to the plate cannot exceed that which just neutralizes the force due to the plate at the filament surface, assuming the initial velocity of the emitted electrons to be negligible. This limitation of the plate current by space charge is the condition which also obtains in the normal operation of a three electrode vacuum tube and which permits the grid to control the flow of plate current. For if the voltage of the grid is positive so that its electrical force aids that of the plate, the current will increase until the combined forces of grid and plate are neutralized at the surface of the filament. If the grid voltage is negative, it will itself neutralize the force of the plate to some extent and the plate current will be reduced. On account of its position in the tube, a change in voltage of the grid will produce a change in the electrical force at the filament, and a resulting change in plate current much greater than an equal change in voltage of the plate. The amplification constant is a measure of the relative effect of grid and plate voltages on the plate current and may be defined as the ratio of changes of the plate and grid voltages necessary to produce equal changes in the plate current.

In the case of the two parallel planes the flow of electrons as limited by the space charge was found by Langmuir to be

$$i_p = 2.33 \times 10^{-6} \frac{V_p^{\frac{3}{2}}}{x^2} \tag{1}$$

in amperes per square centimeter of surface. In terms of the electric force due to the plate,

$$f = \frac{V_p}{x}$$

the current is

$$i_p = 2.33 \times 10^{-6} \frac{f^{\frac{3}{2}}}{x^{\frac{1}{2}}} \tag{2}$$

This equation may be interpreted to mean that the current i_p flowing from filament to plate exerts a force at the filament equal and opposite to the force f. If f is increased or diminished the current will vary in proportion to the three-halves power of the force. If the electrical force f arises from the joint action of both a plate and a grid, it will be assumed that the plate current will vary in accordance with this same law. In case the grid becomes positive enough so that an appreciable flow of electrons to the grid takes place, this can be taken account of by substituting the total current $i_p + i_g$ in place of i_p. These assumpcause the density of the space charge is borhood of the filament where the at the lowest velocity. This makes the pend mainly upon the distribution in the ament and the grid.

an equation for the current flow which action of the grid, it is necessary to get an combined electrical forces due to the grid and ements are at certain definite potentials with ament. Maxwell[4] in considering the screening ing of parallel wires has given a solution of the blem of two infinite planes with a grid of equally cal wires between them. This permits an ex- the first case of the two-electrode tube as given by that of a three-electrode tube.

Maxwell's treatment it is found that the force at the one of the planes, which we will consider to be the fila- given by

$$f=\frac{1}{x+\frac{b_1 b_2}{\alpha}}\left(V_p+\frac{b_2}{\alpha}V_g\right) \tag{3}$$

where

b_1 = distance filament to grid.

b_2 = distance grid to plate.

$x = b_1 + b_2$

and

$$\alpha=-\frac{a}{2\pi}\log_\epsilon\left(2\sin\frac{\pi c}{a}\right) \tag{4}$$

where

c = radius of grid wire

a = distance between grid wires.

In the derivation of this equation it is assumed that c is small with respect to a. Substituting equation (3) in (2) we obtain for the equation of the plate current in a three-electrode tube

$$i_p=\frac{2.33\times10^{-6}}{x^{\frac{1}{2}}\left(x+\frac{b_1 b_2}{\alpha}\right)^{\frac{3}{2}}}\left(V_p+\frac{b_2}{\alpha}V_g\right)^{\frac{3}{2}} \tag{5}$$

Equation (5) is of the same form as the expression

$$i_p=A\,(V_p+k\,V_g)^{\frac{3}{2}} \tag{6}$$

which Langmuir has given as representing the characteristic surface of the plate current. In this formula k is the amplification constant. The formula for the amplification constant is therefore

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[4] Maxwell, "Electricity and Magnetism," volume 1, page 310.

$$k = \frac{b_2}{a} \tag{7}$$

The quantity a is given by equation (4) and depends only upon the dimensions of the grid. Hence the amplification constant is independent of the distance from the filament to the grid but is proportional to the distance b_2 from grid to plate.

The internal resistance of the tube in the plate circuit R_p is the reciprocal of the slope of the plate current-plate voltage characteristic. Differentiating (5) with respect to V_p.

$$\frac{\partial i_p}{\partial V_p} = \frac{3}{2} \cdot \frac{2.33 \times 10^{-6}}{x^{\frac{1}{2}} \left(x + \frac{b_1 b_2}{a} \right)^{\frac{3}{2}}} \left(V_p + \frac{b_2}{a} V_g \right)^{\frac{1}{2}} \tag{8}$$

The reciprocal of this expression gives the internal resistance R_p. Hence

$$R_p = \frac{x^{\frac{1}{2}} \left(x + \frac{b_1 b_2}{a} \right)^{\frac{3}{2}}}{3.5 \times 10^{-6} \,.\, S} \frac{1}{\left(V_p + \frac{b_2}{a} V_g \right)^{\frac{1}{2}}} \tag{9}$$

Since the current i_p is the current per square centimeter of area, the total area S must be introduced in the denominator of the expression for the total resistance R_p.

It is of interest to see with what accuracy the formulas (7) and (9) can be applied to calculate the constants of actual tubes from the dimensions, since actual tubes deviate considerably from the simplifying assumptions necessary in the theoretical treatment. Using formula (7), the amplification constant was calculated for a considerable number of types of tubes from dimensions obtained either by measurement of burned out tubes of the same type, from data supplied by the manufacturer or from somewhat difficult measurements upon good tubes. These values of the amplification constant were then compared with those obtained from the electrical measurement of either the same tubes or a number of tubes of the same type. In the case of tubes having plane grids and plates, the agreement between the measured and computed values was very satisfactory over a wide range of tube constructions. The formula when applied to tubes having cylindrical or elliptical plates and grids gives much less satisfactory agreement. The results of the comparison for tubes having plane elements are given in Table 1. This table gives the maker, type of tube, the dimensions used in the calculation and, in the last two columns, the computed and measured values of the amplification constant. Three types of tubes having cylindrical plates and grids gave computed values respectively 20, 40, and 40 per cent. higher than the measured values. The agreement in the case of the 20 per cent value was aided without doubt by the relatively heavy wires used in supporting the grid helix which would increase the measured value, so that in general the computed values would be about 40 per cent high for tubes of this type.

The curves in Figure 1 permit the amplification constant to be readily determined when the diameter of the grid wire in mils (1 mil = 0.001 inch = 0.025 mm.), the distance between the grid wires in millimeters (or turns per inch) and the distance from grid to plate in millimeters, are given. Each curve corresponds to a particular size of grid wire which is customarily drawn to diameters given in mils. The abscissas are given in tenths of a millimeter distance between grid wires (the heavy vertical lines corresponding to turns per inch as numbered) while the ordinates give the values of amplification constant for each two millimeters of distance from grid to plate. Thus the type D tube of Table 1 has 8 mil grid wire spaced 10.7-tenths of a mm. apart. From the curve the amplification constant is 22 for each two mm. distance plate to grid. This distance is 4.8 mm., hence the amplification constant is 2.4 times 22 = 53.

In applying formula (9) to compute the internal plate circuit resistance of actual tubes in which the filament is a wire instead of a plane as assumed in the derivation of the formula, the difficulty arises as to what values should be used for the area. A suggestion as to a possible solution of this difficulty is gained from the second two-electrode problem for which a solution was given by Langmuir.[5] He finds in the case of a straight round wire for the filament and a concentric cylinder as the plate, that the current flow is independent of the wire radius if this is small compared to the radius of the plate but does vary in proportion to the length of the system. It therefore seems probable that in the case under consideration the resistance might vary inversely as the length of the filament so that an equivalent breadth could be found which would be constant for all of the tubes and which, when multiplied by the length of the filament, would give the correct value of the area to use in the formulas.

To test this assumption values of the resistance for unit area

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[5] Langmuir, previous citation.

TABLE 1

Maker	Type	b_2 cm.	a cm.	c cm.	Amplification Constant	
					Computed	Measured
W. E. Co.	D–Voltage Amplifier	0.48	0.107	0.010	52	40–49
W. E. Co.	V–Voltage Amplifier	0.417	0.121	0.009	30	28–29
G. E. Co.	P–High Power	0.435	0.083	0.0038	26	25
De Forest	3.75-inch–High Power	0.6	0.114	0.0032	19	17
De Forest	(VT–21)–Receiving	0.255	0.137	0.010	14	10–12
W. E. Co.	E (VT-2)–Low Power	0.465	0.29	0.010	6.9	6.7–7.7
W. E. Co.	J(VT-1)–Receiving	0.175	0.25	0.020	6.4	5.3–6.8.
W. E. Co.	L-Current Amplifier	0.49	0.35	0.009	4.8	4.5–5.2.

were computed for all of the types of tubes given in Table 1. These values were then divided by the experimentally determined values of the actual tube resistance giving the effective area for each type of tube. Again dividing these values for the area by the length of the filament the equivalent breadth was obtained. The range in the observed resistance values for different tubes of the same type is usually considerable, so that similar variations in the equivalent breadth are to be expected. The results obtained indicated that the equivalent breadth is a constant for the different types of tubes as accurately as tubes of the same type are reproduced. This is shown in Table 2 where the first five columns give the type of tube, the values of x, b_1, and V_p and the computed resistance for a unit area. In the sixth column is given the range in values of the experimentally observed resistances which were obtained at the plate voltage V_p and grid voltage zero. In the last four columns are given the effective area, the filament length and the range and average of the computed values of the equivalent breadth. Apparently an equivalent breadth of 0.5 cm. (0.2 inch) will give computed values of the plate circuit resistance for the different types of tubes which will fall close to or within the range of values observed for a number of tubes of the same type.

Using this value for the equivalent breadth, formula (9) can be rewritten so as to express the internal plate circuit resistance of a tube in terms of its filament length L and the other dimensions. Thus

$$R_p = \frac{x^{\frac{1}{2}}\left(x+\frac{b_1 b_2}{a}\right)^{\frac{3}{2}}}{1.8L} \frac{10^6}{\left(V_p+\frac{b_2}{a}V_g\right)^{\frac{1}{2}}} \text{ ohms} \qquad (10)$$

This formula is applicable only to tubes having a plane plate and plane grid on each side of the filament.

In conclusion it might be well to call attention to certain experimentally observed deviations in the behavior of actual tubes from that corresponding to the simple theory. At low plate voltages or low values of (V_p+kV_g) the measured values of the amplification constant are lower and the three-halves power law does not appear to hold. This is probably a result of neglecting the initial velocity of emission, the voltage gradient along the filament, and so on. For positive grid voltages when appreciable current flows to the grid, the measured and effective values of amplification constant and internal plate resistance are affected

TABLE 11

Type	x cm.	b_1 cm.	V_p volts	Resist. in 10^5 ohms		Effective Area cm^2.	Filament Length cm.	Equiv. Breadth	
				Computed Per $\frac{1}{\text{cm.}^2}$	Meas'd Total			Range cm.	Average cm.
D	0.65	0.17	100	6.71	1.13–3.25	5.9–2.1	5.7	1.04–0.36	0.70
V	0.62	0.20	100	1.21	0.49–0.51	2.5–2.4	5.7	0.44–0.42	0.43
P	0.64	0.22	300	2.11	0.22	9.6	16.	0.60	0.60
3.75″	0.80	0.20	300	1.45	0.27	5.3	10.	0.53	0.53
VT–21	0.36	0.10	20	0.89	0.68–1.21	1.3–0.74	2.4	0.55–0.31	0.43
E	0.63	0.16	200	0.36_5	0.035–0.065	10.6–5.6	13.	0.81–0.43	0.62
J	0.35	0.15	20	0.67	0.16–0.32	4.2–2.1	5.7	0.74–0.37	0.55
L	0.71	0.21	100	0.17_2	0.049–0.052	3.5–3.3	5.7	0.62–0.58	0.60

thereby. The quantities k and R_p as given above are no longer determined by the slopes of the plate current characteristics alone but by the slopes of the grid characteristics also. Thus if we write as mentioned above

$$i_p+i_g=A\ (V_p+k\,V_g)^{\frac{3}{2}}$$

then

$$\frac{\partial\,i_p}{\partial\,V_g}+\frac{\partial\,i_g}{\partial\,V_g}=\frac{3}{2}A\,k\,(V_p+k\,V_g)^{\frac{1}{2}}$$

and

$$\frac{\partial\,i_p}{\partial\,V_p}+\frac{\partial\,i_g}{\partial\,V_p}=\frac{3}{2}A\ (V_p+k\,V_g)^{\frac{1}{2}}$$

from which

$$k=\frac{\dfrac{\partial\,i_p}{\partial\,V_g}+\dfrac{\partial\,i_g}{\partial\,V_g}}{\dfrac{\partial\,i_p}{\partial\,V_p}+\dfrac{\partial\,i_g}{\partial\,V_p}}=\text{constant}$$

$$R_p=\frac{1}{\dfrac{\partial\,i_p}{\partial V_p}+\dfrac{\partial\,i_g}{\partial V_p}}$$

while the measured and effective values would be given by

$$(k)_{meas.}=\frac{\dfrac{\partial\,i_p}{\partial\,V_g}}{\dfrac{i_p}{\partial\,V_p}}=\text{function of }V_g$$

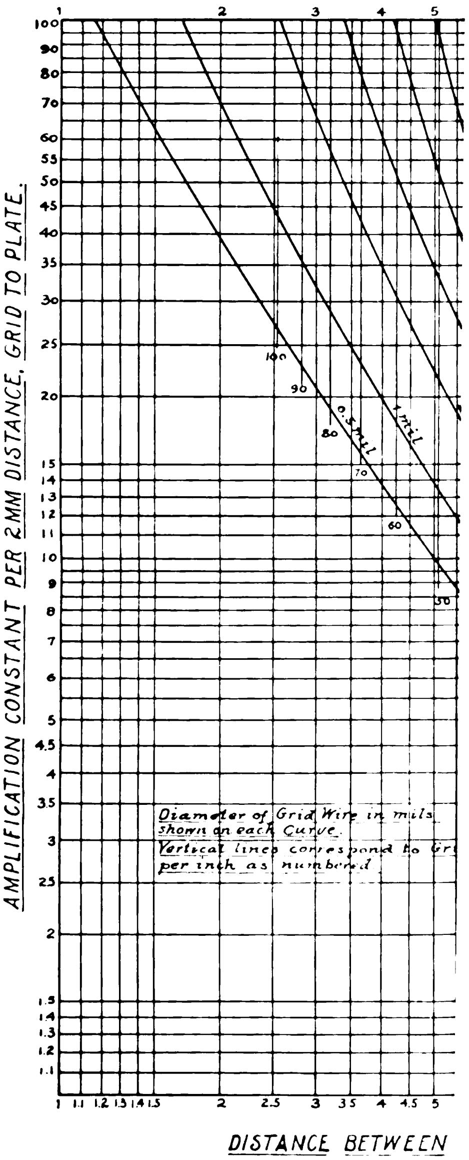
$$(R_p)_{meas.}=\frac{1}{\dfrac{\partial\,i_p}{\partial V_p}}$$

It is not to be expected therefore that the formulas (7) and (9) will give results in close accord with the measurements when (V_p+kV_g) is small or V_g is considerably positive.

In conclusion, I beg to acknowledge the valuable assistance of Dora E. Wells of the Bureau of Standards, who performed most of the experimental work cited above.

SUMMARY: The amplification constant of three electrode vacuum tubes is mathematically derived in terms of readily measurable physical dimensions of the tube. Curves are plotted enabling the rapid pre-determination of tube amplification constant on this basis. The curves are shown to give results agreeing with experiment for a number of types of tubes.

The internal resistance of the tube is similarly deduced, and an agreement between theory and experiment shown. Certain deviations in specific instances are explained.


AMPLIFICATION CONSTANT PER 2MM DISTANCE, GRID TO PLATE.
DISTANCE BETWEEN
1
2
3
4
5
100
90
80
70
60
55
50
45
40
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1.5
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3
3.5
4
4.5
5
100
90
80
70
60
50
0.5 mil
1 mil
Diameter of Grid Wire in mils shown on each Curve.
Vertical lines correspond to Gr
per inch as numbered

AN EXPERIMENT ON IMPULSE EXCITATION*

BY

JOHN H. MORECROFT

(ASSOCIATE PROFESSOR IN ELECTRICAL ENGINEERING, COLUMBIA UNIVERSITY, NEW YORK CITY)

The question of how much current will flow in a given circuit, when it is excited by impulse excitation, is capable of mathematical analysis by the application of Fourier's Integral, or other methods, providing the shape of the exciting pulse is given. However, an experimental analysis is always worth while as a verification of the theoretical conclusions.

Atmospherics (strays or static) are undoubtedly a kind of pulse; certain experiments seem to show that these atmospheric pulses are sometimes oscillatory, with rather high damping, but the probability is that most of the disturbance from static is due to unidirectional pulses impressed on the antenna by sudden motions of electricity, either in the atmosphere or in the earth. With the idea of showing how a unidirectional pulse affects an oscillatory circuit, an experiment was carried out with the circuit arranged as shown in Figure 1.

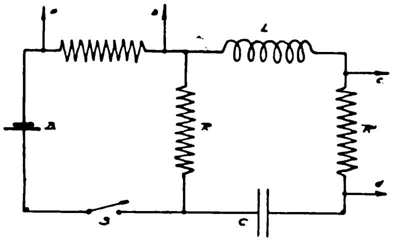

FIGURE 1

The oscillating circuit was made up of a coil of 0.206 henry and condenser of 20 microfarads, having a natural frequency of 78.6 cycles per second. A battery *B* was arranged to cause

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* Received by the Editor, April 12, 1919.

current to flow thru the resistance R when the switch S was closed. This switch was so built that its time of contact could be varied from about 0.002 second to any longer time desired. The resistance of R was 0.01 ohm.

Another resistance was put in the circuit at R', the value of this being about 5 ohms. The resistance R was made low so that the current thru it from the battery would be very much larger than the current in the oscillatory circuit, to keep the current from the battery constant whatever the form of the current in the oscillatory circuit. The current from the battery was 10 amperes while the current in the oscillatory circuit was only a fraction of an ampere.

The voltage impressed in the oscillating circuit by the battery is the IR drop thru the resistance R and the form and duration of this current was obtained by one vibrator of the oscillograph being connected across the points a-b. Another vibrator connected across the terminals of the resistance R' gave the form of the current in the oscillatory circuit.

Films were taken for various times of contact of the switch S, from the shortest time possible to somewhat greater than the natural period of the circuit. Several of the films are given in Figures 2-5; they show the form of pulse used for excitation and also the magnitude of the resulting current. It is also possible to obtain at once from the film the ratio of the pulse duration to the natural period of the circuit. For the longer pulses

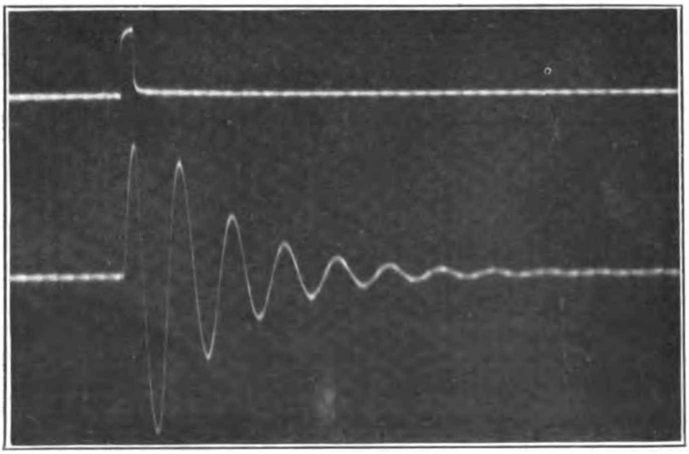

FIGURE 2

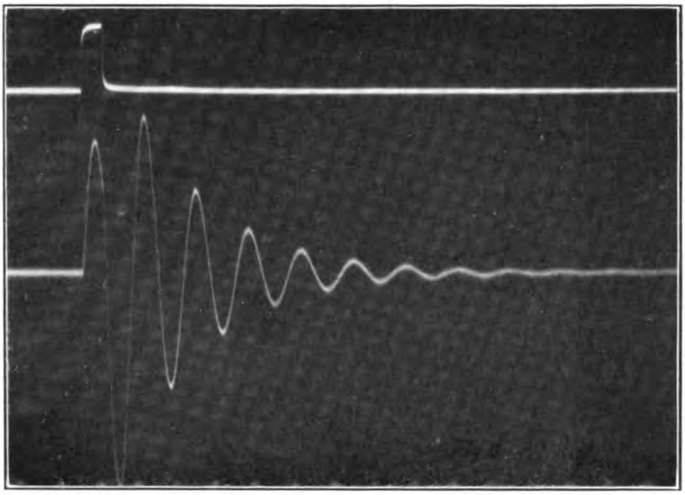

FIGURE 3

the form approximates a rectangle, but the shortest ones (the corners of the pulse being absent) have a form more like sine wave.

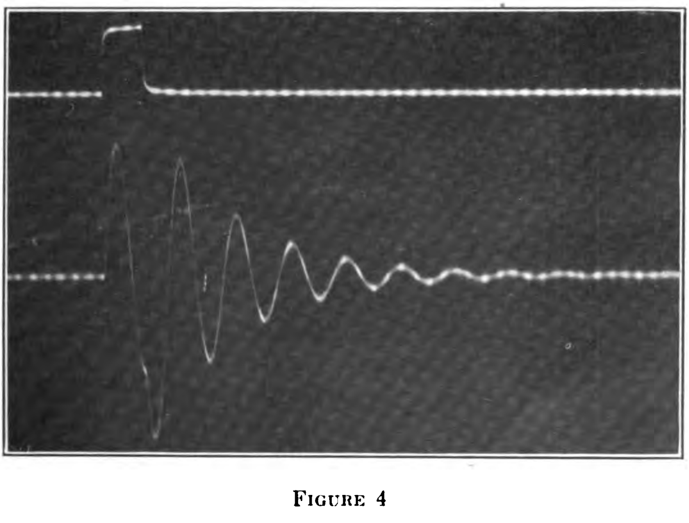

FIGURE 4

The amplitudes of the first and second alternations of the oscillating current were measured and these values plotted in the form of a curve, using as abscissas the ratio of the pulse length to the natural period of the circuit. These values are given in Figure 6. It is seen that for the shortest time used the amplitude of the first and second alternations are practically the same and that the amplitude of the first alternation does not increase after the pulse has a length equal to one quarter of a cycle. As the pulse was lengthened the amplitude of the second alternation increased until the pulse length was equal to practically one-half the natural period of the oscillating circuit. If the

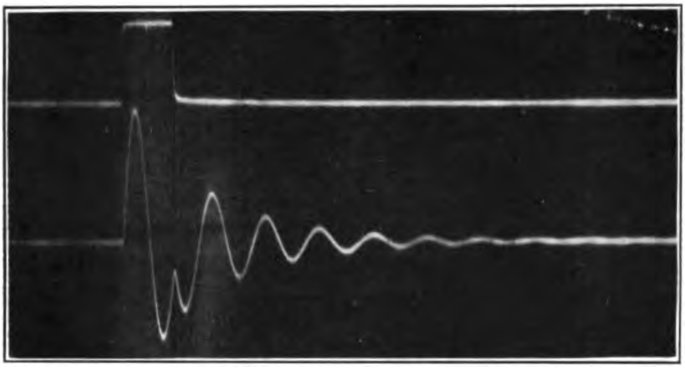

FIGURE 5

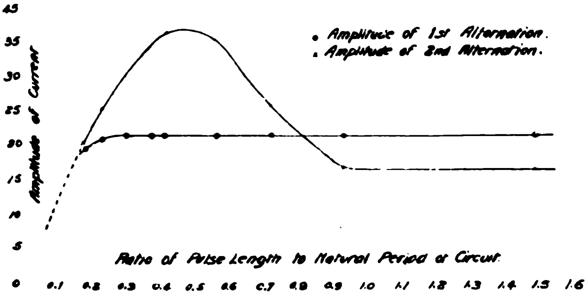

FIGURE 6

decrement of the circuit had been lower the greatest amplitude would have been produced by a pulse somewhat longer than that which did actually give the greatest amplitude in the test.

For pulses lasting longer than one-half period the amplitude of the second alternation decreases until it has that value which is fixed by the amplitude of the first alternation and the decrement of the circuit; for any longer pulse the amplitude of the first and second alternations are not affected.

It is seen therefore that the greatest disturbance is produced in an oscillatory circuit by a rectangular pulse when this pulse has a duration equal to one-half the natural period of the circuit.

The complex form of current occurring in an oscillatory circuit as a result of irregularly timed pulses, of various durations, is shown in Figures 7 and 8; if this form of excitation occurs at very short intervals of time, the frequency of the oscillating circuit being above audibility, the resulting noise heard in the phones of a coupled circuit (such as the ordinary closed circuit of a receiving circuit) would be the scratchy noise known as static.

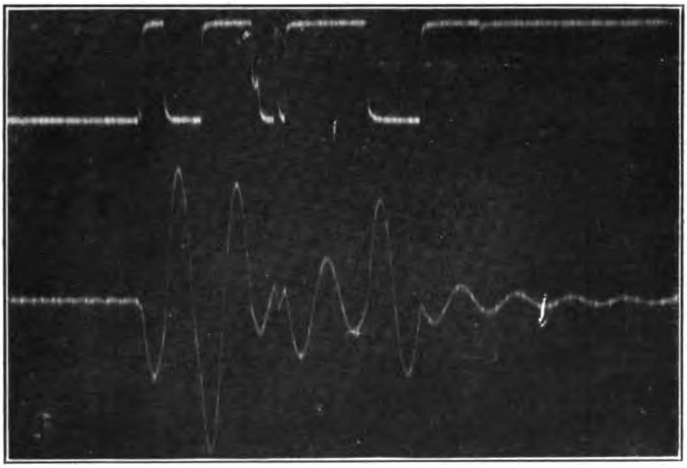

FIGURE 7

It is to be noted that the pulse of emf. shown in Figures 3-8 was introduced directly into the oscillating circuit. I mention this point because incorrect conclusions might be drawn if it [illegible] suggested that the results were applicable to shock excitation in which a pulse of current in one circuit is utilized to excite [illegible] another magnetically coupled to it.

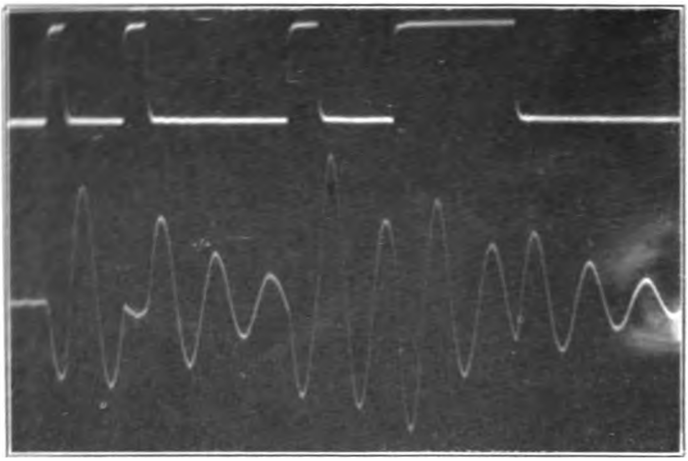

FIGURE 8

If a square pulse of current of form similar to the emf. pulse of Figures 3-8) is allowed to flow in a primary circuit it will not generate oscillations in the coupled circuit in agreement with the conclusions reached above. This is due to the fact that a square current pulse in one circuit will not generate a square emf. pulse in a coupled circuit, but a very different form of pulse. Figure 9 shows about what would occur; a trapezoidal form of current in the primary circuit would induce in the secondary circuit two very short pulses of emf., these two pulses being in opposite directions. Hence the action of such a current pulse in the primary, in so far as producing current in the secondary is concerned, is the same as tho two very short pulses of emf. were introduced directly into the secondary circuit, the two pulses being in opposite directions and separated by a time about equal to the length of current pulse in the primary circuit.

Since the above notes on impulse excitation were written the author attended a meeting of THE INSTITUTE OF RADIO ENGINEERS at which another phase of impulse excitation was brought up and on which there is apparently some difference of opinion.

If a so-called infinite impedance circuit is used in an antenna for weeding out a certain frequency, how well will this circuit weed out pulses, or, what amounts to the same thing, how well will it prevent interference from a damped wave station? Suppose an inductance, *L*, and a condenser, *C*, are connected in parallel as indicated in Figure 10; then if the antenna is excited by a signal of the same frequency as that for which the parallel

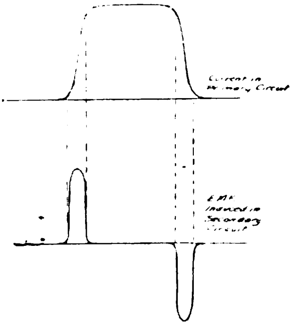

FIGURE 9

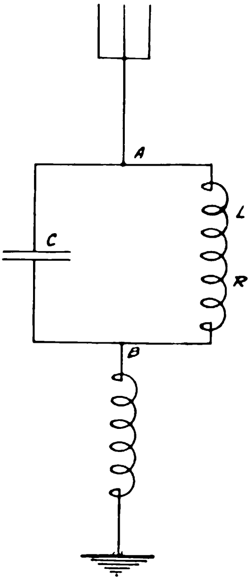

FIGURE 10

circuit is tuned, it is shown in many text books that the impedance offered by this parallel path is resistance only and its value is equal to the actual resistance of the coil multiplied by the square of the ratio of the coil reactance to the coil resistance, this on the assumption that the condenser resistance is negligible. This impedance may be thousands of times as much as the resistance of the coil; thus if the coil has an actual resistance of 5 ohms and a reactance-to-resistance ratio of 200, (a quite possible figure), the impedance between the points *A-B* is 200,000 ohms, *for the steady state of an alternating current of the proper frequency.*

Now the question arises, will this circuit act towards a pulse with such a high impedance, and of course the answer is *No.* Suppose the critical frequency of the parallel circuit is 100,000 cycles and to make the analysis as simple as possible let an emf. pulse be impressed on the antenna, the pulse being sinusoidal in form and lasting 0.000,005 second. This assumption makes the pulse the same shape as one alternation of that emf. for which the parallel path offers such a high impedance; it seems evident that such a pulse will produce less current thru the circuit than any other form. If the pulse is not sinusoidal in form it may be resolved (according to a quite generally accepted idea) by a Fourier's analysis, and the different members of the series obtained may be treated separately. Thus the sinusoidal pulse assumed may be considered as the result of such an analysis of a more complex pulse.

The current flowing in the antenna as the result of excitation by such a shaped pulse is shown in Figure 11. The first curve (a) is the emf. impressed, the second curve (b) is the current thru the condenser, and the third curve (c) is the current in the inductance branch, this being drawn on the assumption that the resistance of the coil is negligible with respect to its reactance. The current flowing in the antenna (or line in which the infinite impedance circuit is introduced) is obtained by adding the two branch currents and is shown by the rectangular current (d) of Figure 11. The amplitude of this rectangular current is the same as the maximum value of the current flowing in the condenser; quite evidently the parallel path permits the flow of a large current and so can scarcely be considered as having infinite impedance, even for a pulse of the form most favorable for the filtering action of the circuit.

The reason that the predicted impedance does not exist for the impulsive emf. is this—the equations from which the value of this infinite impedance are obtained neglect the transient terms and for impulsive forces these transient terms are the whole solution.

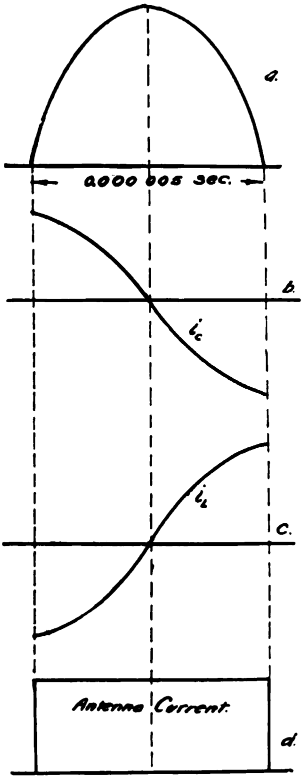

FIGURE 11

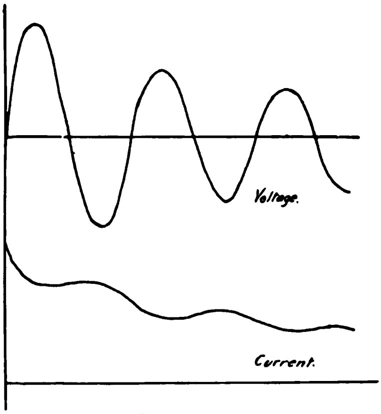

FIGURE 12

In case the circuit of Figure 10 is relied upon to weed out interference from a spark station it must be remembered that at the beginning of every wave train there are transient terms which do not obey the laws holding for the steady state. In Figure 12 is shown what may be expected when the infinite impedance circuit is excited by a damped wave of emf. The current which flows in the antenna (in so far as it is limited by the infinite impedance circuit) is shown to be a quite appreciable current, rising at once to the maximum current in the condenser branch and then gradually dying away with slight ripples of the interfering signal frequency. Such a current flowing in the antenna coil of the receiving circuit would certainly be very difficult to tune out.

SUMMARY: There is described an oscillographic investigation of the current produced in an oscillatory circuit by a brief rectangular pulse of emf. directly introduced into the circuit. The pulse length is regulable. It is found that a pulse having a duration of one-half the natural period of the circuit produces the maximum disturbance. The bearing of the results on the effects of strays in reception is discussed.

The behavior of "infinite impedance" loop circuits when subjected to pulse excitation is studied, and it is found that for pulses their filtering action is much reduced.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 8 APRIL, 1920 Number 2

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THE COLLEGE OF THE CITY OF NEW YORK

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NAVAL AIRCRAFT RADIO*

BY

T. JOHNSON, JR.

(EXPERT RADIO AID, NAVY DEPARTMENT, WASHINGTON, D. C.)

Another low power equipment, developed and manufactured by the General Electric Company, is known as type SE 1340. This was designed for short range continuous wave buzzer-modulated telegraph and telephone transmission from small flying boats such as type HS-2-L. It consists essentially of a vacuum tube continuous wave transmitter of approximately 5 watts antenna input. No provision is made for telegraph transmission by continuous wave. Transmission on wave lengths of 335, 375, and 425 meters is arranged for, the following normal operating ranges being obtained:

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilometers	Nautical miles	Kilometers
Telephone	60	111	20	37
Buzzer-modulated	75	139	30	55.5

The communication ranges tabulated in this paper do not represent the maximum obtainable, but merely the dependable operating range in daily service. Perfect telephone communication has been conducted with this set on a trailing antenna from aircraft to shore over a distance considerably in excess of the range here given, but daily operation of a large number of these sets under service conditions, has proven that this is a dependable operating range.

The total weight of this equipment installed is 110 pounds (50 kilograms). The transmitter may be operated intermittently on one standard 12-volt battery for four hours.

The main element of this equipment, as illustrated in Figure

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*This is the second half of Mr. Johnson's paper. It is continued from the February, 1920, PROCEEDINGS.—EDITOR.

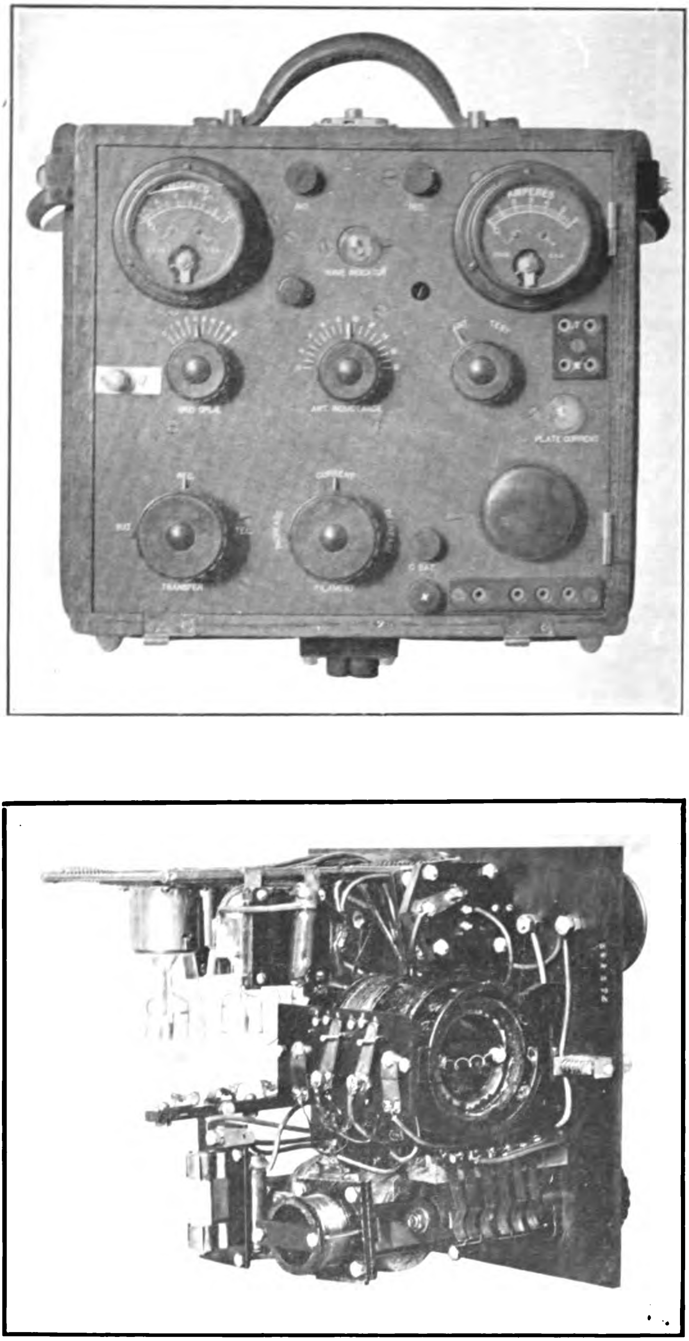

AMPERES
AMPERES
TUNE INDICATOR
ANT. INDUCTANCE
PLATE CURRENT
REC.
CURRENT
TRANSFER
FILAMENT
C BAT.

FIGURE 62

62, is a transmitting cabinet 11 inches (28 cm.) wide by 10 inches (25.4 cm.) high by 8.5 inches (21.6 cm.) deep, and weighing 11 pounds (5 kilograms). On the panel front of the transmitter cabinet are mounted a combination send-receive and signal switch, filament rheostat, buzzer, grid coupling control, knob for varying antenna inductances to compensate for the slight antenna variation on various planes, artificial-antenna switch with spring release to prevent the artificial antenna from being left in circuit, wave meter lamp, wave meter button, modulation indicator lamp, filament and antenna ammeters, and terminals for connecting leads. The send-receive switch starts the dynamotor. On the back of the panel are mounted the inductances, condensers, switch mechanisms, tubes on spring mountings, and other elements of the apparatus, the panel being hinged so that the entire apparatus may be swung out for inspection and repair without removing the set from its installed position or disconnecting the leads. The panel is mounted on slip hinges so that it may be entirely removed without dismounting the container cabinet.

The schematic plan of connections is shown in Figure 63. Three transmitting tubes, type CG 1162, are utilized, one as an oscillator and two as modulators. Power for the set is supplied from a standard 12-volt battery, 5.25 amperes being required for the tube filaments and 10 amperes for the dynamotor. The dynamotor, illustrated in Figure 64, is 8 inches

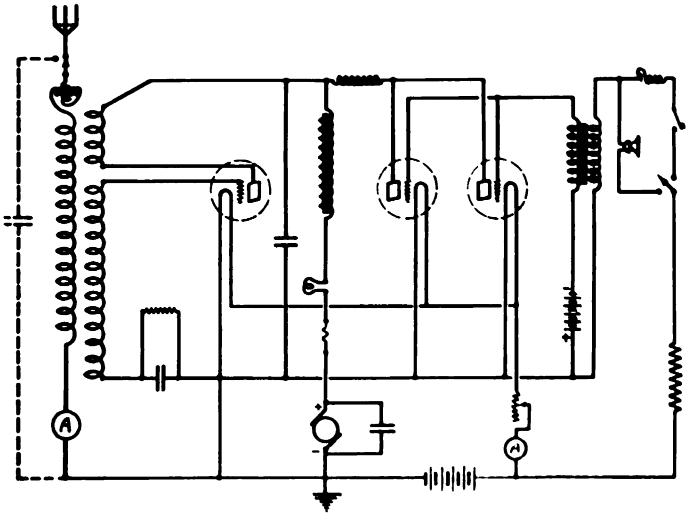


FIGURE 63

(20.3 cm.) long by 6 inches (15.2 cm.) wide by 6.5 inches (16.5 cm.) high, and weighs 11 pounds (5 kilograms). It supplies 65 milliamperes at 350 volts to the plate circuits of the tubes.

This set was originally arranged for use of two small negative grid-potential batteries mounted on the back of the panel. Later developments made possible the elimination of this battery.

The constant endeavor to reduce the size and weight of equipment resulted in an improved design of this set. Due to constant development in improving the operating characteristics of the tube used, it became possible to accomplish the same results as in the above set by the use of a single modulator tube. Other improvements made it possible to reduce the size of the transmitter cabinet to 8.25 inches (20.9 cm.) wide by 8.5 inches

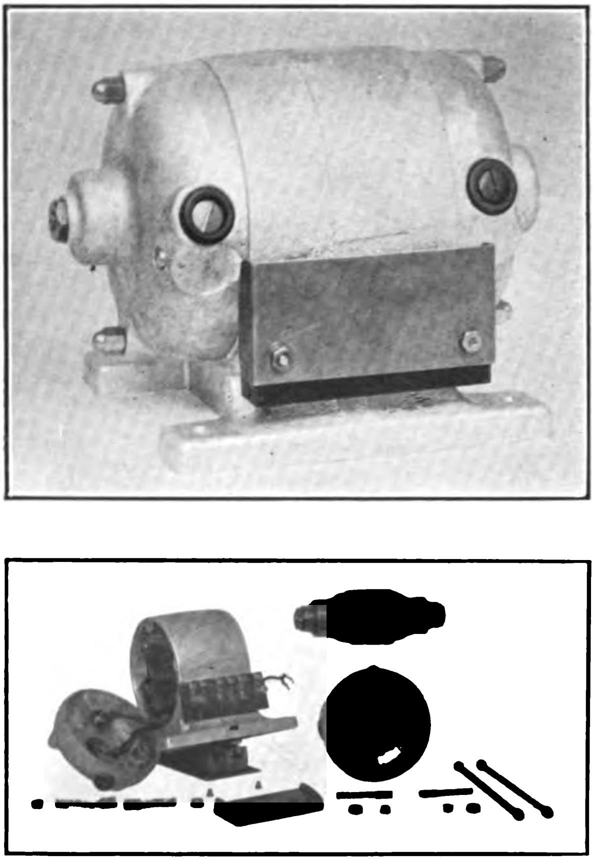
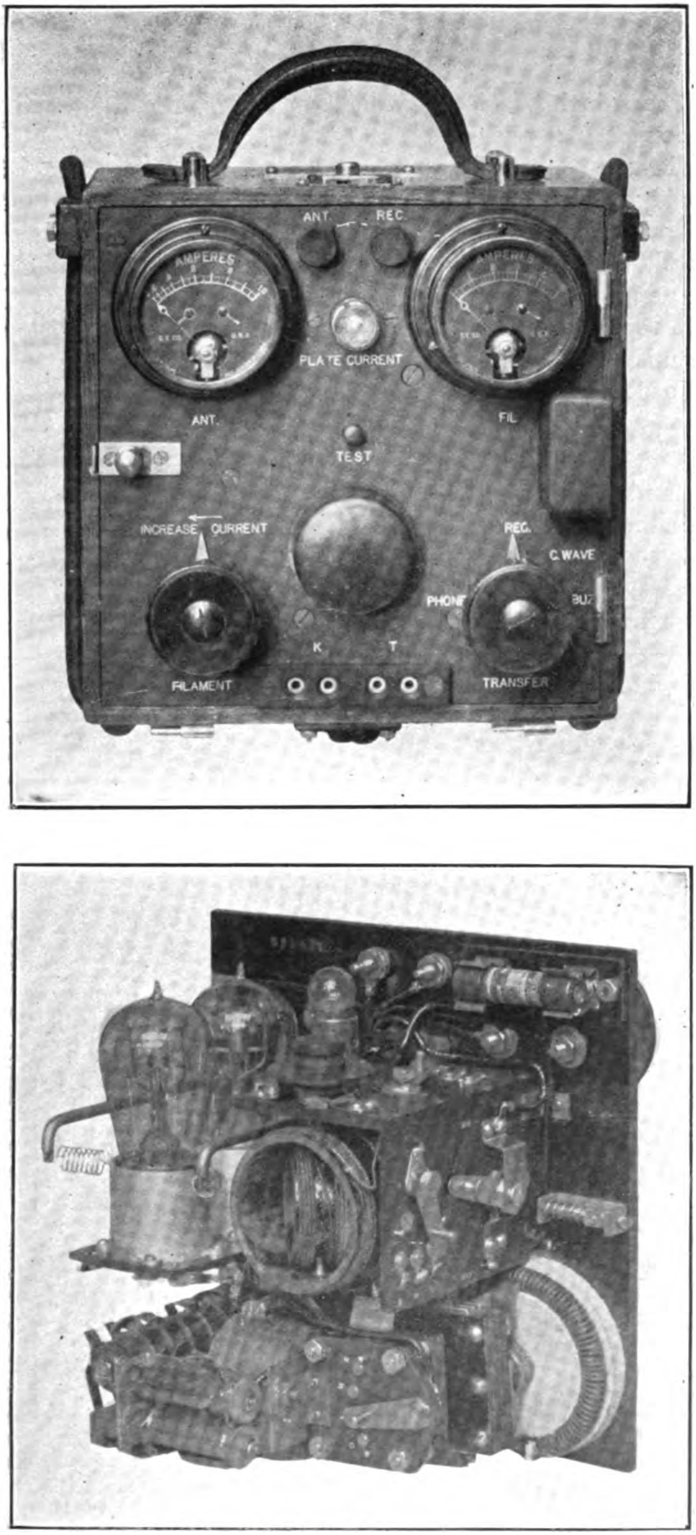


FIGURE 64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REC.
AMPERES
AMPERES
PLATE CURRENT
ANT.
FIL
TEST
INCREASE CURRENT
REG.
C.WAVE
PHONE
BUZ
K
T
FILAMENT
TRANSFER

FIGURE 65

(21.6 cm.) high by 5.5 inches (14 cm.) deep, and the weight to 8.5 pounds (3.9 kilograms).

The main element of this equipment is the transmitting cabinet shown in Figure 65. On the panel front are mounted a combination send-receive and signal switch, filament rheostat control, buzzer, filament and antenna ammeters, wave meter lamp, and terminals for connecting leads.

In this set is added the feature of continuous wave telegraph transmission, the following normal operating ranges being obtained:

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilo-meters	Nautical miles	Kilo-meters
Telegraph	100	185	50	92.5
Telephone	60	111	20	37
Buzzer-modulated	75	139	30	55.5

The schematic plan of connections is shown in Figure 66 Power for this set is also supplied from a standard 12-volt battery, 3.5 amperes being required for the tube filaments, and 9

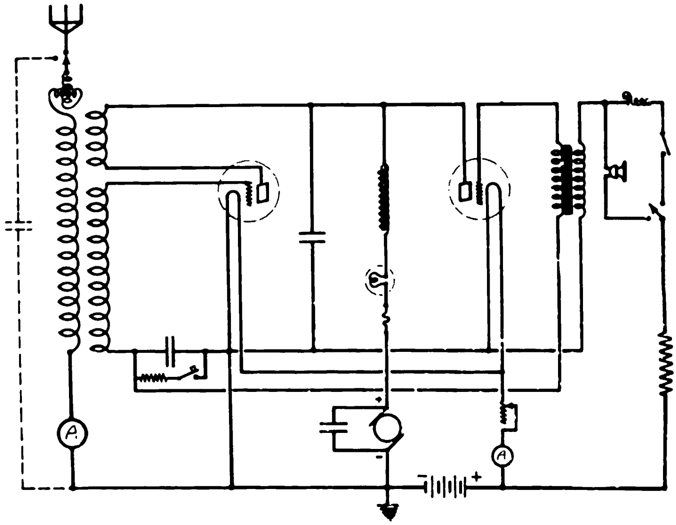


FIGURE 66

amperes for the dynamotor. The dynamotor, similar to that supplied with the previously described set, supplies 55 milliamperes at 350 volts to the plate circuits of the tubes. The transmitter may be operated on one standard 12-volt battery for four hours and thirty minutes.

This transmitter was used on the type NC flying boats on the trans-Atlantic flight, and is shown in the installation illustrated in Figure 101.

A transmitter of power similar to the two equipments just described is type CAG 1295, manufactured by the General Radio Company, and illustrated in Figure 67. This transmitter utilizes two tubes, type CG 1162, as oscillators and one as a modulator. The results obtained with this transmitter have been far inferior to those obtained with the two sets just described, and it is therefore not considered as a standard equipment.

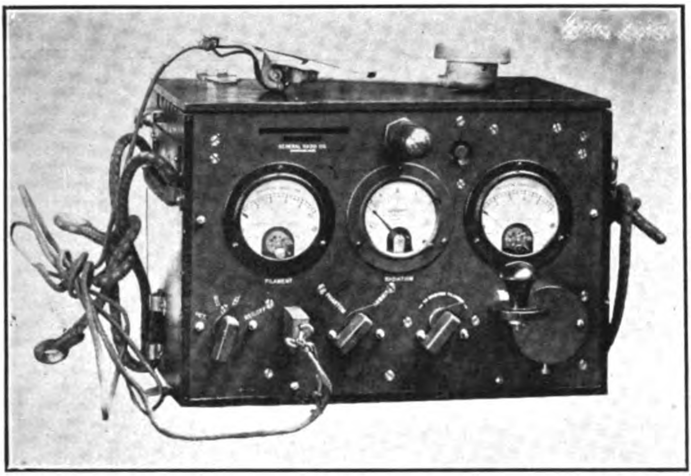


FIGURE 67

Equipment type SE 1370, developed and manufactured by the General Electric Company, was designed for medium range continuous wave telegraph and telephone transmission from large flying boats such as type F-5-L. It consists essentially of a vacuum tube continuous wave transmitter of approximately 100 watts antenna input in the case of continuous wave telegraph and 50 watts in the case of telephone transmission. Provision is also made for buzzer-modulated signalling. Transmission on wave lengths of 600 meters and 425 or 850 meters is arranged for the following operating ranges being obtained:

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilo-meters	Nautical miles	Kilo-meters
Telegraph	175	324	90	167
Telephone	85	157	45	83.3
Buzzer-modulated	100	185	60	111

The transmitters are normally supplied for 600 and 425 meters with an additional 850-meter coil system which may be used to replace the 425-meter coil system.

The main element of the equipment illustrated in Figure 68 is a transmitter 15 inches (38 cm.) wide by 15 inches (38 cm.) high by 8.25 inches (21 cm.) deep, weighing 17.5 pounds (7.93 kilograms). The dynamotor for this equipment, illustrated in Figure 69, is 6 inches (15.2 cm.) wide by 7 inches (17.8 cm.) long by 7.75 inches (19.7 cm.) high, and weighs 17 pounds (7.7 kilograms).

The schematic diagram of connections is shown in Figures 70 and 71. Two vacuum tubes, type CG 1144, are used, one as an oscillator and one as a modulator. Power is supplied from a 24-volt battery, consisting of two standard 12-volt units, 6.5 amperes at 24 volts being required for the tube filaments and 15 amperes at 24 volts being required to operate the dynamotor. The dynamotor supplies 300 milliamperes at 750 volts to the plate circuits of the tubes for continuous wave telegraph signalling, and 200 milliamperes for modulated transmission.

Deriving power from a 24-volt battery consisting of two standard 12-volt units in series, this transmitter may be operated 1.5 hours under service conditions.

The next higher power equipment of this general type is type SE 1380, developed and manufactured by the General Electric Company. This was also designed for use on large flying boats such as type F-5-L and on dirigibles. It consists essentially of a vacuum tube continuous wave transmitter of approximately 250 watts antenna input for continuous wave telegraph transmission and 150 watts antenna input for modulated transmission. In the design of this set the antenna input

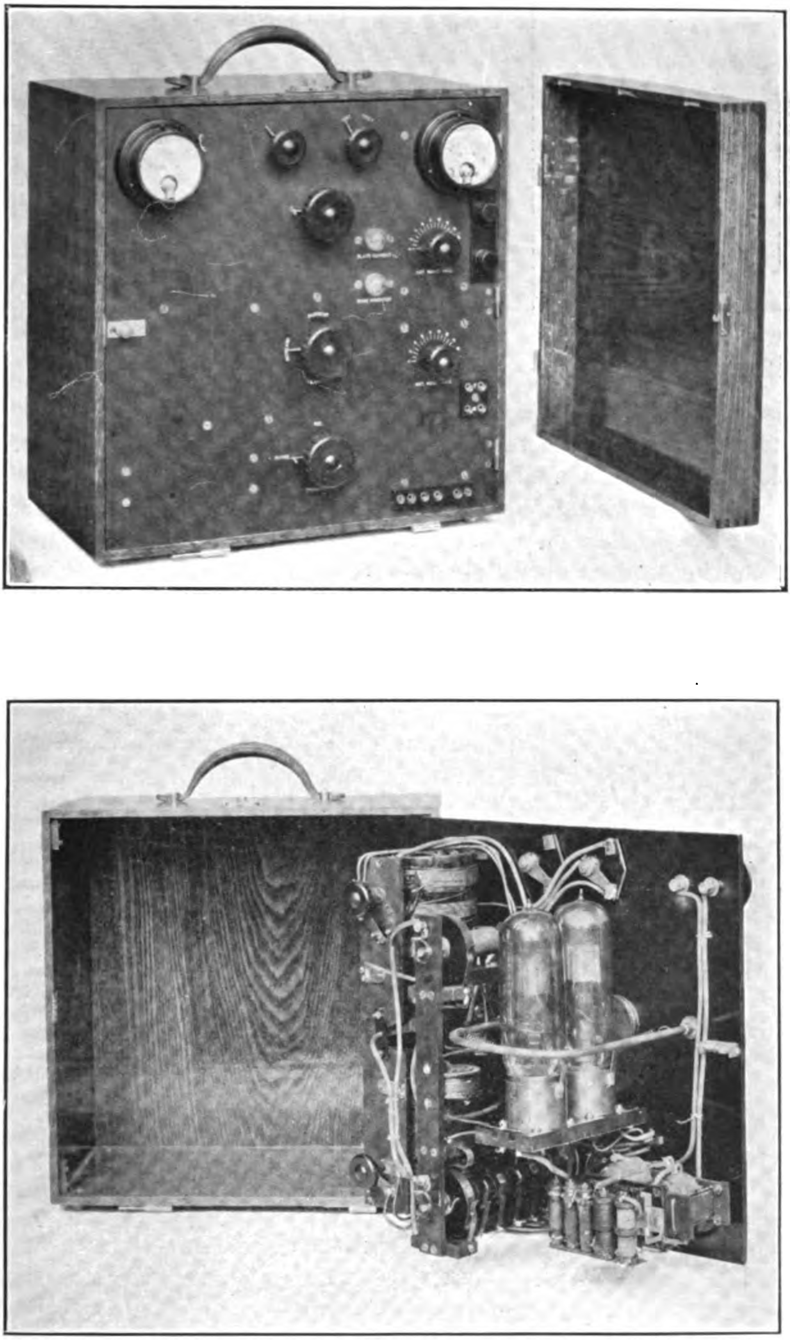


FIGURE 68

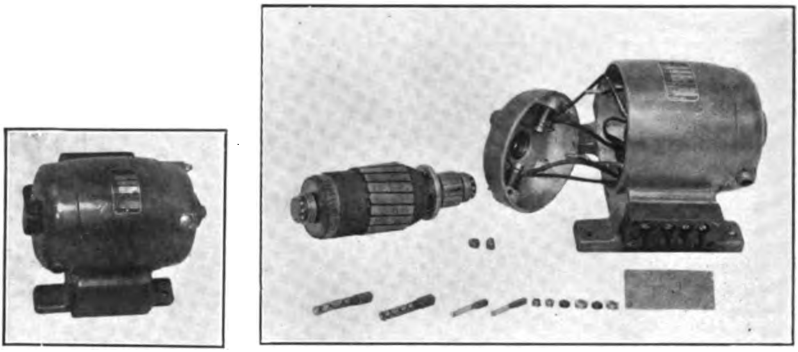


FIGURE 69

is limited only by the generator power limitations, which are influenced by considerations of weight of generator and battery. Provision is made for buzzer-modulated and telephone signalling in addition to the continuous wave telegraph. Transmission on wave lengths of 600 and 1,600 meters is arranged for, the following normal operating ranges being obtained:

FOR 600 METERS

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilometers	Nautical miles	Kilometers
Telegraph	250	463	175	324
Telephone	150	278	75	139
Buzzer-modulated	190	352	90	167

FOR 1,600 METERS

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilometers	Nautical miles	Kilometers
Telegraph	160	296	135	250
Telephone	125	232	100	185
Buzzer-modulated	140	260	115	213

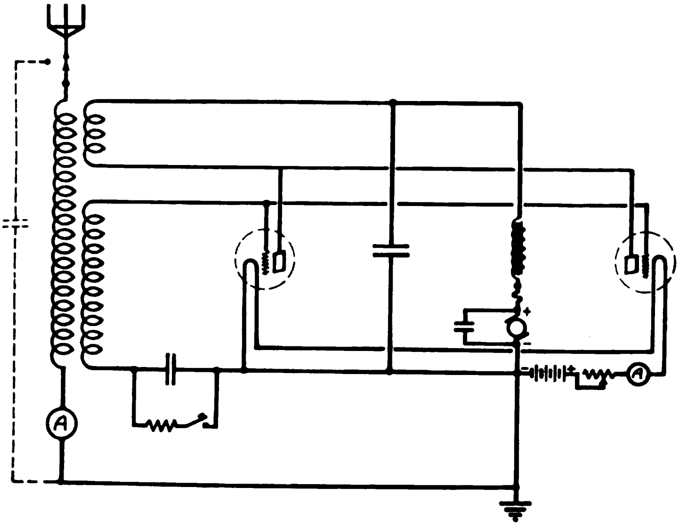


Figure 70

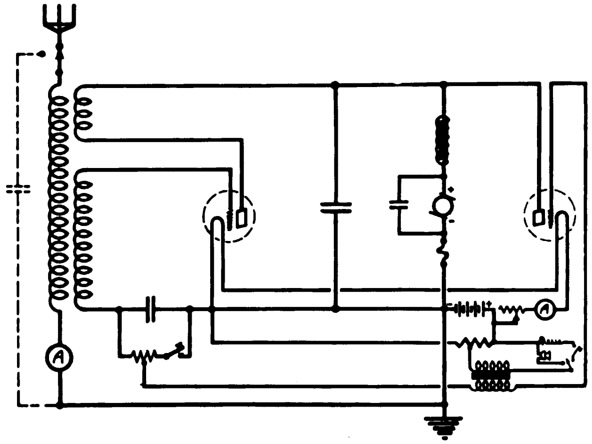


Figure 71

The main element of the equipment, illustrated in Figure 72, is a transmitter cabinet 20 inches (50.8 cm.) wide by 19 inches (49.5 cm.) high by 11.5 inches (29.2 cm.) deep, and weighing 40 pounds (18.1 kilograms). The dynamotor for this equipment with protective condenser, illustrated in Figure 73, is 6.5 inches (16.5 cm.) wide by 11.5 inches (29.2 cm.) long by 7 inches (17.8 cm.) high, and weighs 29 pounds (13.1 kilograms). The

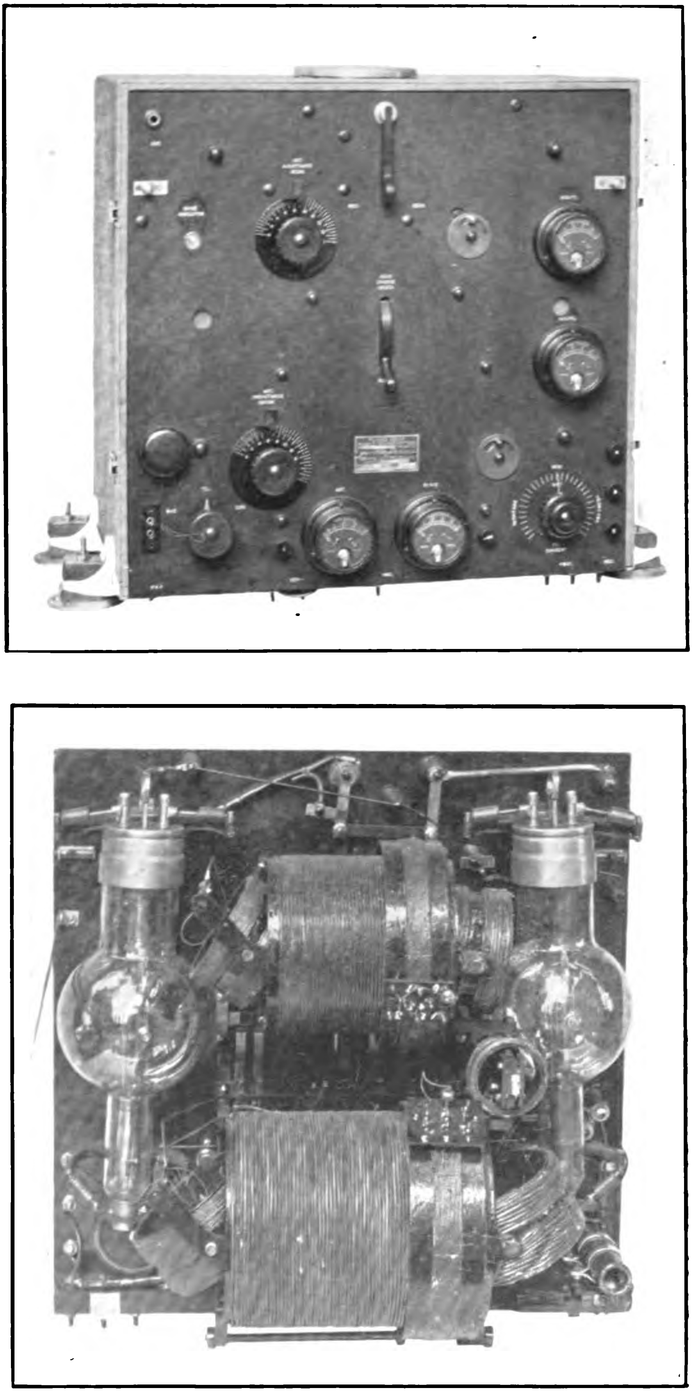


FIGURE 72

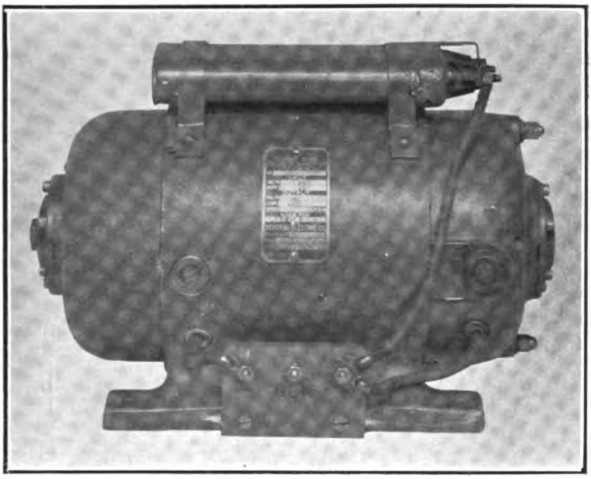


Figure 73

starting switch for the dynamotor and fuse for the high tension circuit are separately mounted.

The schematic diagram of connections is shown in Figure 74. Two vacuum tubes, type CG 916, are used, one as an oscillator and one as a modulator. Power is supplied from a 24-volt battery consisting of two 12-volt standard units, 3.6 amperes

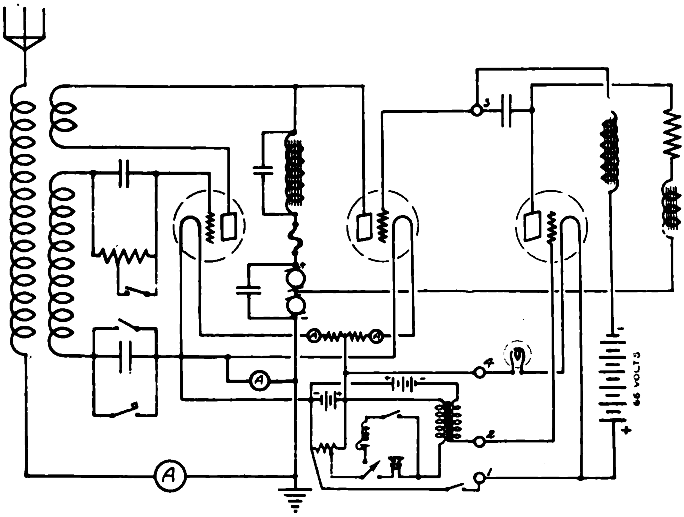


Figure 74

at 24 volts being required for the tube filaments and 18 amperes at 24 volts being required to operate the dynamotor. The dynamotor supplies 320 milliamperes at 1,500 volts to the plate circuits of the tubes for all methods of transmission. In this set the modulation is indicated by the fluctuation of the plate circuit ammeter, and in addition, by the wave meter lamp.

In the design of this set, maximum efficiency is attained by using two entirely separate coil systems for the two wave lengths utilized. For these coils there was developed a special litzendraht wire consisting of four cables of thirty-two 0.005-inch (0.127 mm.) wires each. Great care was necessary in the design of this set due to the high potentials built up as the result of operating on a plate voltage of 1,500. The use of the heavy vacuum tubes involved the design of special tube mountings, using concentric springs, an outer heavy spring absorbing heavy shocks due to landing, and an inner lighter spring absorbing the normal aircraft vibration.

Deriving power from a 24-volt battery consisting of two standard 12-volt units in series, this transmitter may be operated one hour under service conditions.

It has been shown that in the radio telephone transmitters developed by the General Electric Company, modulation of the radio frequency energy is effected by imposing the audio frequency output of the microphone transformer on the grid circuit of the modulator tube. The amplified audio frequency output of the modulator tube is then introduced into the plate circuit of the oscillator tube.

It has been found that in sets where a considerable amount of radio frequency energy is to be modulated, such as transmitters utilizing tubes types CG 916 or CG 1144 as oscillators, better results may be obtained by using a modulation amplifier. This amplifier, consisting of a single stage of amplification, is inserted between the output side of the microphone transformer and the input, or grid circuit, of the modulator tube. Its function is to amplify the output of the microphone transformer which in turn results in an increased output from the modulator tube.

The schematic circuit diagram of the modulation amplifier is shown in Figure 75 in which "*F*," "*G*," and "*P*" represent the filament, grid and plate of a tube type CG 1162. "*R*" is a high resistance and "*L*" and "L_1" are iron core reactances. "*T*" is the microphone transformer and "*C*" a condenser. The batteries "*B*" and "B_1" are used to maintain the proper grid potential on the amplifier and modulator tubes. The 750-volt power is ordinarily supplied from the same generator or dynamotor used to supply the plate circuits of the oscillator and

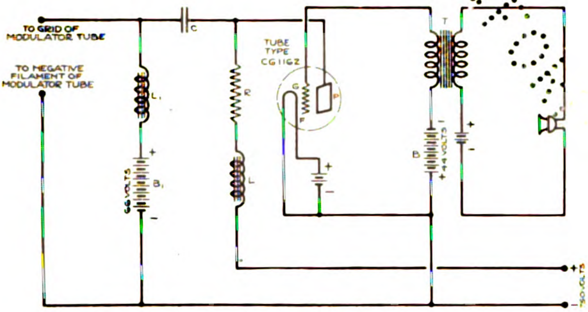

TO GRID OF MODULATOR TUBE
TO NEGATIVE FILAMENT OF MODULATOR TUBE
C
L₁
+
66 VOLTS
B₁
−
R
L
TUBE TYPE CG1162
G
P
F
+
−
T
B
+44 VOLTS
+
−
+
−
750 VOLTS

FIGURE 75

modulator tubes. The modulation amplifier, arranged in a small cabinet and known as type CG 4030, is illustrated in Figure 76.

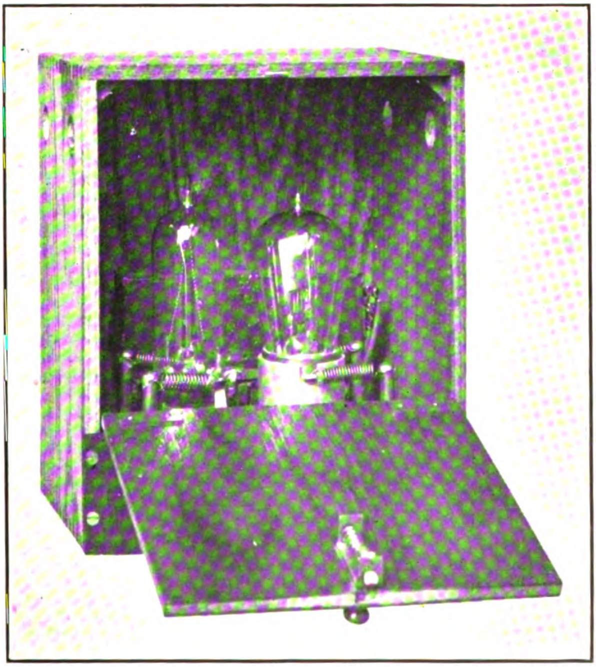


FIGURE 76

Another high power equipment utilizing two large tubes, type CG 916, is that known as type SE 1100, developed and manufactured by the Marconi Wireless Telegraph Company. This set is also designed for use on large flying boats such as type F-5-L and on dirigibles. Altho utilizing the same dynamotor and the same tubes as equipment type SE 1380, this set uses less power. This fact, together with a lower efficiency due to energy consumed in the set itself, results in an antenna input of only 100 watts for continuous wave telegraph transmission and 60 watts antenna input for modulated transmission. Transmission on wave lengths of 600 and 1,600 meters is arranged for, the following normal operating ranges being obtained:

For 600 Meters

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilo-meters	Nautical miles	Kilo-meters
Telegraph	150	278	70	130
Telephone	60	111	35	65
Buzzer-modulated	80	148	45	83

For 1,600 Meters

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (In Water)	
	Nautical miles	Kilo-meters	Nautical miles	Kilo-meters
Telegraph	90	167	50	93
Telephone	40	74	25	46
Buzzer-modulated	45	83	35	65

The main element of the equipment, illustrated in Figure 77, is a transmitter panel 19 inches (48.2 cm.) wide by 19 inches

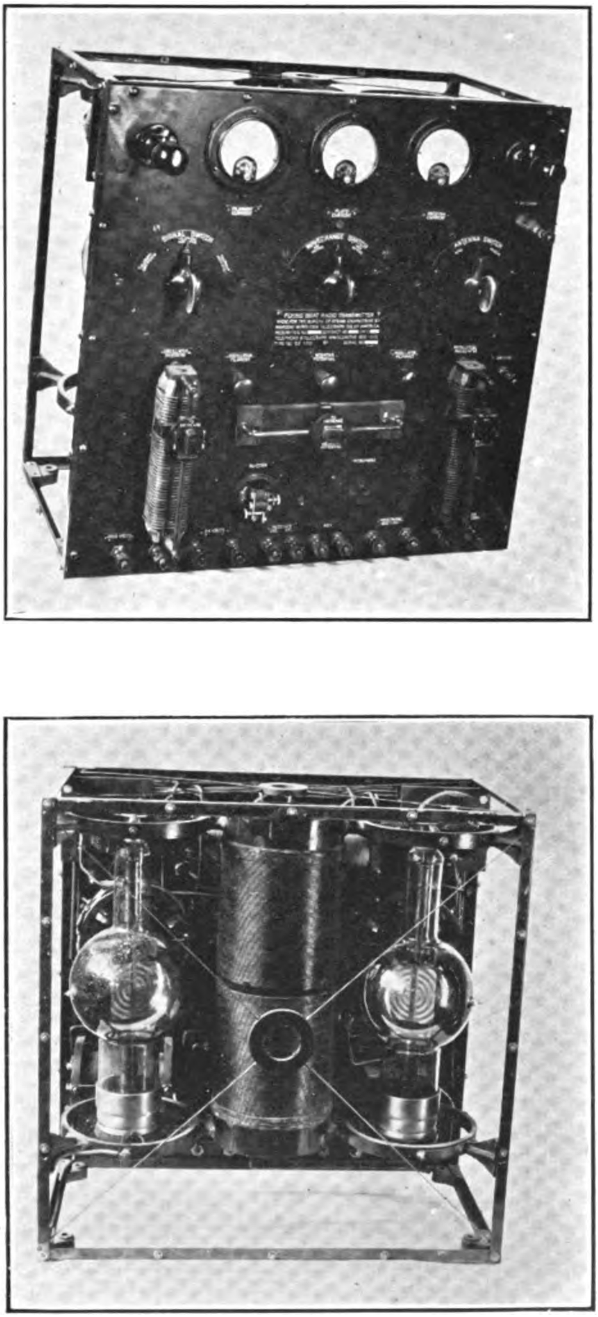

SIGNAL SWITCH
WAVE CHANGE SWITCH
ANTENNA SWITCH
FLYING BOAT RADIO TRANSMITTER

Figure 77

(48.2 cm.) high by 11.5 inches (29.2 cm.) deep, weighing 40 pounds (18.1 kilograms). Plate circuit power at 1,500 volts is supplied from the dynamotor illustrated in Figure 73, 200 milliamperes being used. As no modulation amplifier is employed, the modulation is quite inferior to that obtained with equipment type SE 1380. The schematic diagram of connections is shown in Figure 78.

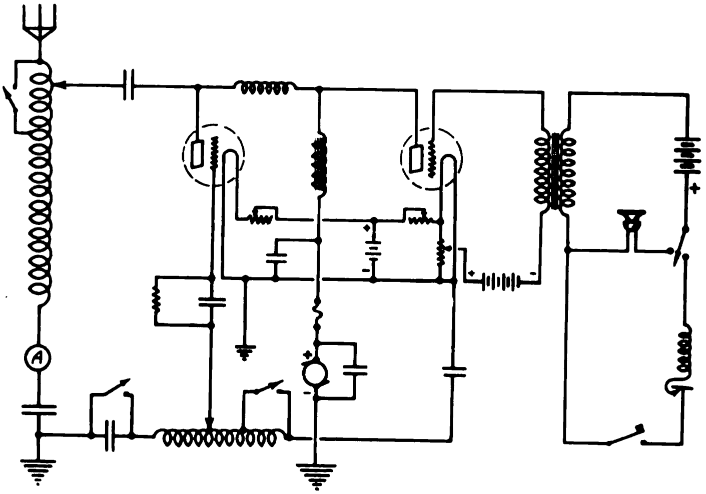


FIGURE 78

The loss of power in the transmitter itself is due to the fact that the set employs a closed oscillation circuit containing several condensers and two coils. Such a circuit naturally consumes considerable radio frequency energy. When changing from 1,600 to 600 meters, an additional loss is incurred by short circuiting a section of both plate and grid coils. The circuit employed was chosen because of its stability of operation, but since the development of this set, this point has proven of little importance, as other circuits employed have given equally stable operation and yet have involved but low internal losses. Another disadvantage of this set is the fact that it is highly critical to antenna adjustment. In reviewing the inferiority of this set with respect to the type SE 1380 equipment, it should be stated that the former was a considerably earlier development. The type SE 1100 equipment was the first of this general type and power placed in quantity production for use during the war and the time available for the completion of the design was very limited.

Figure 79 illustrates a standard installation of an SE 1100 equipment together with a standard receiver on a type F-5-L flying boat.

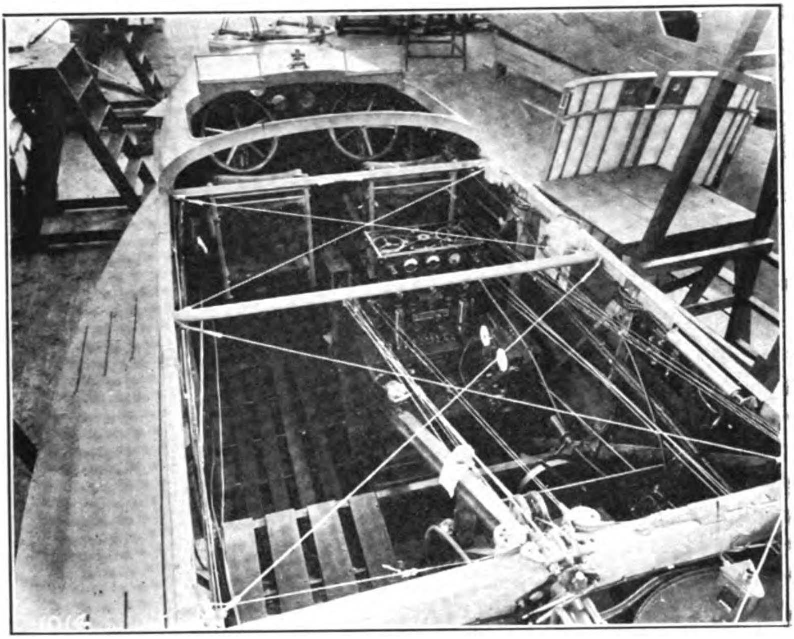


FIGURE 79

The highest power transmitter of the vacuum tube type used on naval aircraft is that known as type SE 1390, developed and manufactured by the General Electric Company. This was designed for use on the largest flying boats, such as type NC, and on large dirigibles. It consists essentially of a vacuum tube continuous wave transmitter of approximately 500 watts antenna input for continuous wave telegraph transmission, and 250 watts antenna input for modulated transmission. Provision is made for buzzer-modulated and telephone signals in addition to the continuous wave telegraph. Transmission on wave lengths of 600 and 850 meters is arranged for, the following normal operating lengths being obtained:

For 600 or 850 Meters

Method of Signalling	Trailing Antenna (In Flight)		Skid Fin Antenna (On Water)	
	Nautical miles	Kilometers	Nautical miles	Kilometers
Telegraph	600	1110	350	648
Telephone	200	370	125	232
Buzzer-modulated	225	417	140	260

The main element of the equipment illustrated in Figure 80 is a transmitter cabinet 20 inches (50.8 cm.) wide by 20 inches (50.8 cm.) high by 16 inches (40.7 cm.) deep, weighing 55 pounds (24.9 kilograms). The dynamotor for this equipment, 7.5 inches (19 cm.) in diameter and 13.19 inches (33.5 cm.) long, weighs 60 pounds (27.2 kilograms).

The schematic diagram of connections for continuous wave telegraph is illustrated in Figure 81; for telephone or buzzer-modulated telegraph in Figure 82. Six vacuum tubes, type CG 1144, are utilized, the entire six as oscillators for continuous wave telegraphy. For telephone or buzzer-modulated transmission three of the tubes are used as oscillators, two as modulators, and one as a modulation amplifier. Power is supplied from a 24-volt source, 20 amperes being required for the tube filaments and 55 amperes to operate the dynamotor. The dynamotor supplies one ampere at 1,000 volts to the plate circuits of the tubes.

The tubes are mounted on a special spring support which, together with the radio coils, switches, wiring, and so on, are all supported from the panel proper which is mounted by means of detachable hinges to the cabinet, so that the panel may readily be moved for inspection and repairs. Special attention has been given in the design to obtain simplicity of wiring and, altho the cabinet is not excessively large, ample room has been allowed for proper mechanical support and insulation. The cabinet, when installed, is mounted on soft rubber supports at the four corners.

The wave length and character of signal is controlled by a single switch and handle which appears on the right side of the

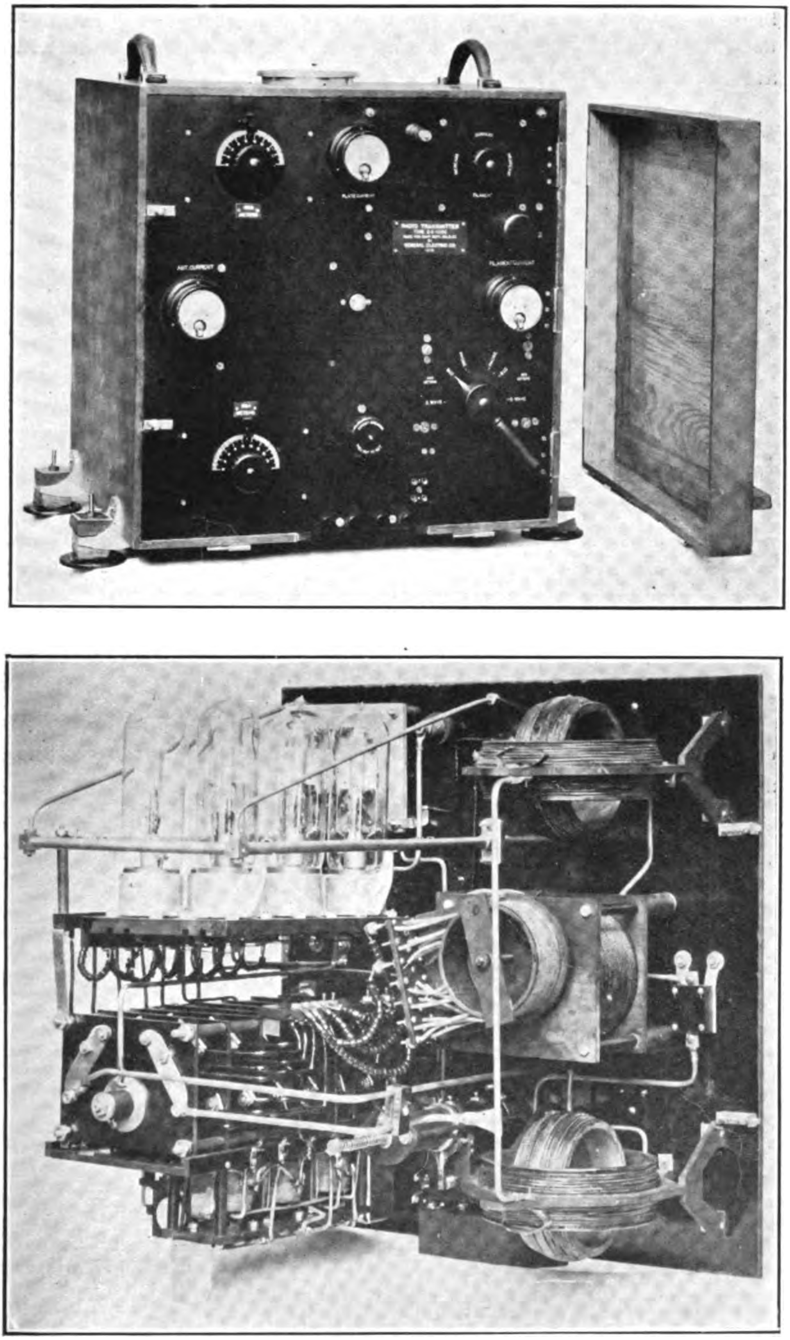

RADIO TRANSMITTER
GENERAL ELECTRIC CO.
PLATE CURRENT
ANT. CURRENT
FILAMENT CURRENT

Figure 80

panel. A special 0.5-ohm rheostat of the multi-step type is used to control the filament current. These two adjustments are the only ones which the operator is required to control.

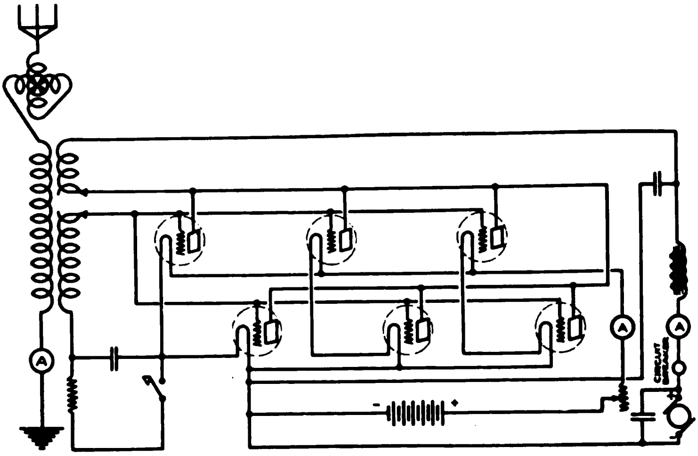


FIGURE 81

A special feature of this set is the one-ampere, 1,000-volt circuit breaker, which is extremely light and rugged, and has a non-resetable feature.

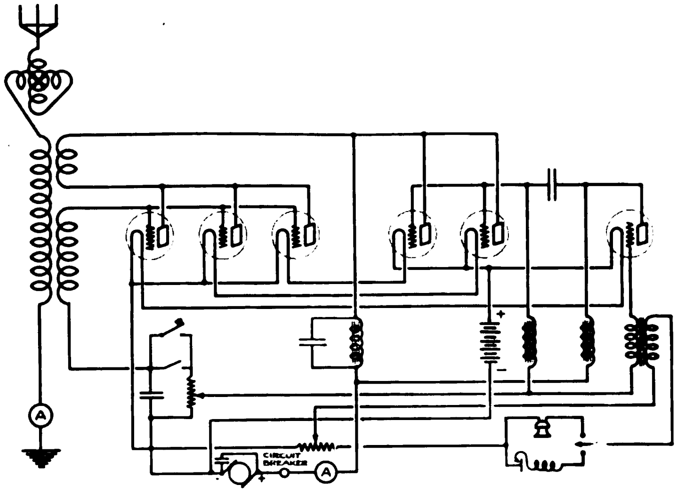


FIGURE 82

STANDARD EQUIPMENT—RECEIVING AND RADIO COMPASS APPARATUS

The reception of radio signals on aircraft is an entirely different problem from that of any other form of radio reception, and early in the aircraft radio work furnished a very novel experience to those who had been accustomed to receiving radio signals under the comparatively highly favorable conditions of a land station. In brief, the difficulties encountered may be classified as acoustic disturbances, consisting of wind rush, engine noise, and vibrational noises; and electrical disturbances resulting from vibration of vacuum tubes and other apparatus, and from induction from engine ignition systems.

The electrical disturbances are provided against by the proper design of receiving apparatus, proper flexible mountings, shielding of ignition systems, and so on. The matter of acoustic disturbances has its solution in the design of a suitable helmet holding the radio telephone receivers. Altho several helmets had been designed for this purpose, none were found satisfactory, and it was necessary to design a new helmet which would be suitable for the needs of the naval service. This helmet, illustrated in Figure 83, is known as type SE 2000. It is made of soft leather with a flannel lining, the central rear seam being left open in manufacture to allow for the fitting of each individual helmet. The main feature of the helmet is the deep soft rubber ear cup which encloses the radio telephone receiver, as illustrated in Figure 84, and fits closely to the head excluding external noises. This may be accomplished by means of a deep cup without placing undue pressure upon the ear itself, a procedure which results in great pain if continued for several hours. This point is of greater importance in the naval service than in that of the land military forces, for in the naval patrols, aircraft pilots and radio operators must wear a helmet for five to eight hours at a time. The depth of the rubber cup is not uniform, but is less back of the ear to accommodate the slightly raised contour of the head at that point. This is very important as continued pressure on this portion of the head results in violent headaches. The rubber cups are held tightly against the head by means of a strap running around the forehead and the back of the neck instead of by a chin strap—a very uncomfortable method. The chin strap is used only to bring the forward edges of the helmet close to the face and to assist in closing the lower portion of the helmet. Another important feature is the flannel lined cape which, when buttoned

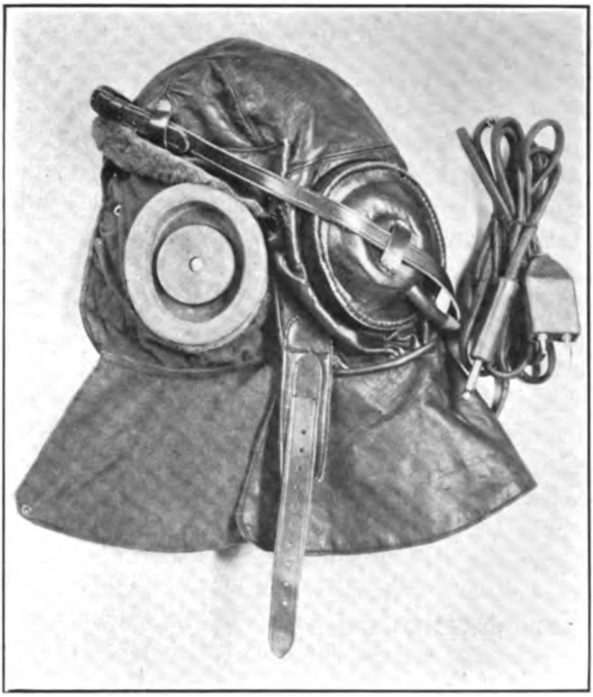
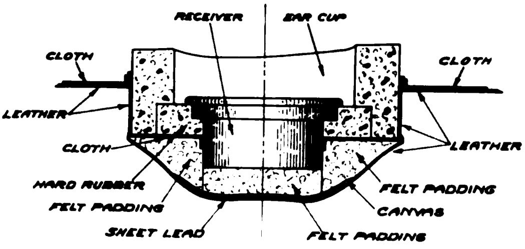


FIGURE 83

within flying clothes, prevents the entry of wind and other noises at this point.

Before describing the standard aircraft receivers it will be of interest to consider briefly the vacuum tubes used in this



— SECTION VIEW —
OF
EAR CUP ON AIRCRAFT RADIO HELMET TYPE SE 2000
SHOWING METHOD OF ASSEMBLY.
FIGURE 84

apparatus. There are at present used for this purpose two types, CW 933 and SE 1444. Type CW 933, illustrated in Figure 85, is a coated filament tube manufactured by the Western Electric Company. Type SE 1444, illustrated in Figure 86, is a tungsten filament tube of Navy Department design, manufactured by Moorhead Laboratories, Incorporated. The latter has nickel elements and is a low filament current tube. It has an amplification constant and a plate resistance somewhat higher than Type CW 933, but its capacity is considerably lower. The elementary data regarding these tubes as used under service conditions is tabulated in Figure 87.

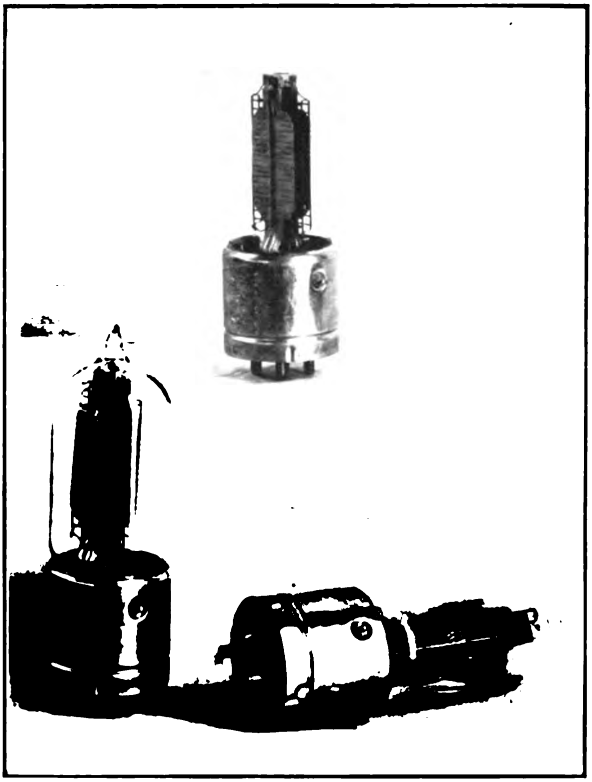


FIGURE 85

The first standard naval aircraft receiver is type SE 950, illustrated in Figure 88. This receiver, mounted in a cabinet

12.63 inches (32 cm.) wide by 13.5 inches (34.3 cm.) high by 5.81 inches (20.6 cm.) deep, and weighing 21 pounds (9.5 kilograms), consists of an inductively coupled vacuum tube receiver of a wave length range from 300 to 2 500 meters provided with static tube coupling for attaining regeneration and oscillation, with two stages of audio frequency amplification. This receiver is also provided with the proper switching and a compensating inductance to adapt it for radio compass use. The schematic diagram of this receiver, omitting the radio compass features, is illustrated in Figure 89. This receiver utilizes three tubes type CW 933. These tubes are mounted on a special damped spiral spring mounting. The entire receiver is not provided with any additional flexible mounting. This receiver was designed at the Radio Test Shop, Navy Yard, Washington, D. C., and has been manufactured by the National Electrical Supply Company.

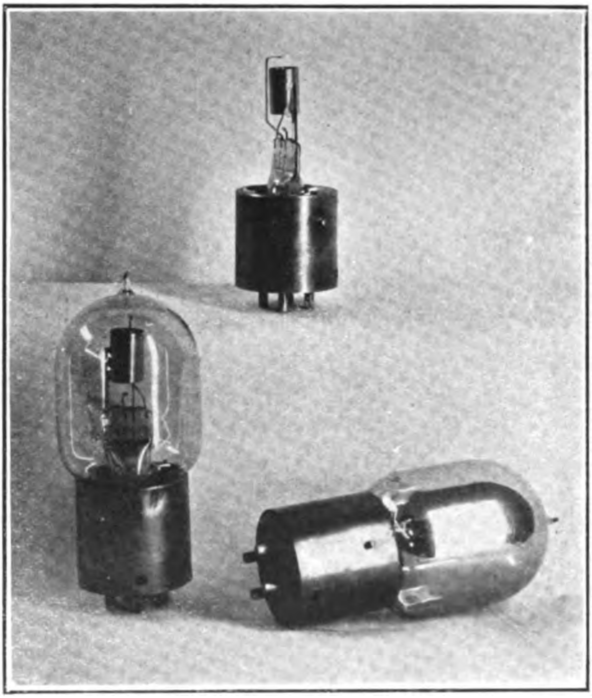


FIGURE 86

Aircraft radio receiver type SE 1414, illustrated in Figure 90, is contained in a cabinet 11 inches (27.9 cm.) wide by 10 inches (25.4 cm.) high by 5.13 inches (13 cm.) deep, and weighs 10.75 pounds (19.9 kilograms). It consists of a conductively coupled receiver of wave length range 200 to 3,000 meters with inductive tube coupling for regeneration and oscillation, and is provided with two stages of audio frequency amplification. In this receiver the tubes are not provided with any individual flexible mounting but the entire receiver is mounted in a rubber suspension shown in the illustration. The schematic circuit diagram of this receiver is shown in Figure 91; three tubes type SE 1444 are used. The lightness and compactness of this receiver results

Navy Type Number	CW 933	SE 1444
Signal Corps Type No.	VT-1	None
Manufacturer	Western Electric Company	Moorhead Laboratories, Incorporated
Manufacturer's Type No.	J	
Filament volts	2.4	4.5
Filament amperes	1.1	0.65
Plate volts	40	40
Grid bias voltage	2.2	1.3
Amplification Constant	5	9
Detector Constant*	$37\frac{(\text{microamperes})}{(\text{volt squared})}$	$20\frac{(\text{microamperes})}{(\text{volt squared})}$
Mutual Conductance	330 micromhos	180 micromhos
Plate Resistance	15,000 ohms	50,000 ohms
Capacity Grid-to-Filament	7.1 micromicrofarads	3.3 micromicrofarads
Capacity Plate-to-Filament	9.2 micromicrofarads	4.2 micromicrofarads
Length over-all	4.06 inches 10.3 cm.	4.06 inches 10.3 cm.
Diameter, Maximum	1.34 inches 3.41 cm.	1.75 inches 4.45 cm.
Weight	2.36 ounces 67 grams	1.81 ounces 51.3 grams

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*The Detector Constant given above is equal to the Second Derivative of the Grid Voltage-Plate Current Characteristic of the tube, and, therefore, in addition to expressing the rectifying properties of the tube, also includes its inherent amplification.

FIGURE 87—Receiving Tubes

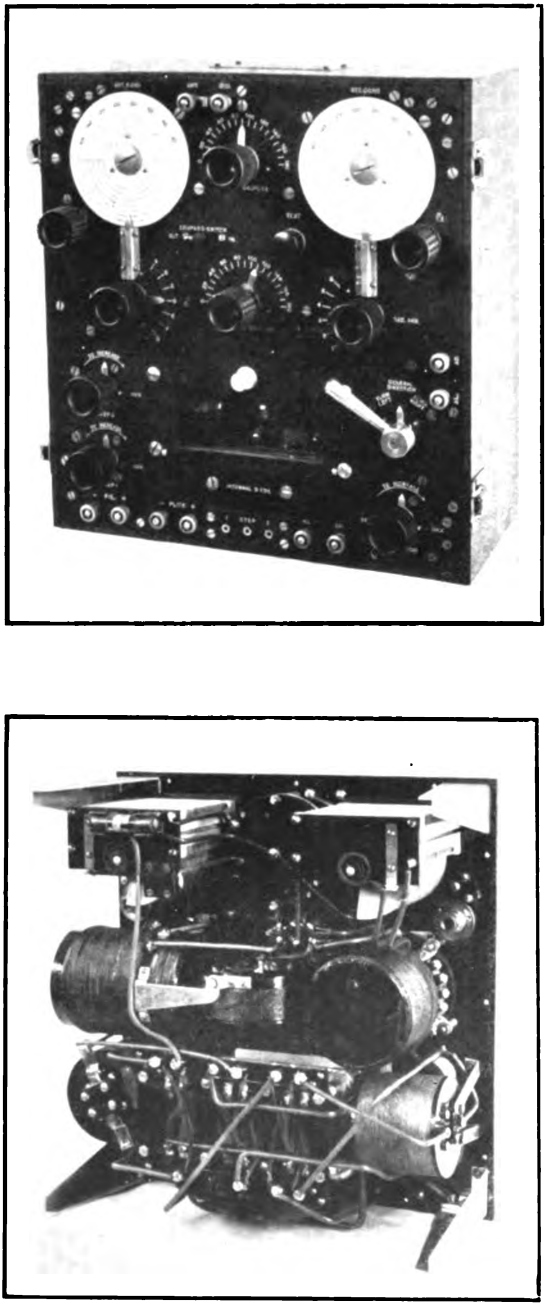

FIGURE 88

largely from the use of small closed iron core transformers, and from small inductance design made possible by the use of special grades of conductor. This receiver was designed at the Radio Test Shop, Navy Yard, Washington, D. C., and has been manufactured by the Westinghouse Electric and Manufacturing Company.

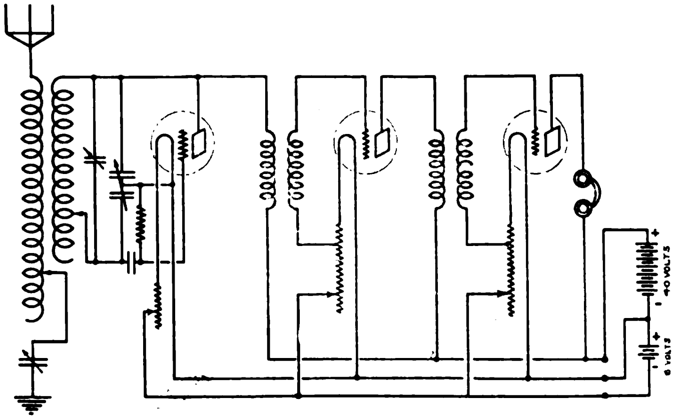

FIGURE 89

Another receiver used in the Naval aircraft service is type CW 1298, illustrated in Figure 92. This receiver merely embodies the receiving portion of telephone equipment type CW 1058, the schematic diagram of which is illustrated in Figure 60. The receiver is 11 inches (27.9 cm.) wide by 9.75 inches (24.7 cm.) high by 4.5 inches (11.4 cm.) deep, and weighs 9 pounds (4.1 kilograms). This receiver is designed and manufactured by the Western Electric Company.

Another very interesting piece of apparatus used for radio reception on naval aircraft is amplifier type SE 1605-B, illustrated in Figure 93. This amplifier is designed primarily for use in connection with radio compass apparatus described later, but is also used successfully for any damped radio reception within the wave length range to which it is adapted. This amplifier is mounted in a cabinet 11.5 inches (29.2 cm.) wide by 8 inches (20.3 cm.) high by 4.06 inches (10.3 cm.) deep, weighing 10 pounds (4.5 kilograms), and consists of three stages of radio frequency amplification, a tube detector, and two stages of audio frequency amplification. The audio frequency amplifying transformers are the small closed iron core type used in receiver type SE 1414. The radio frequency amplifying transformers are of the air core type, and are extremely compact. The transformers are reson-

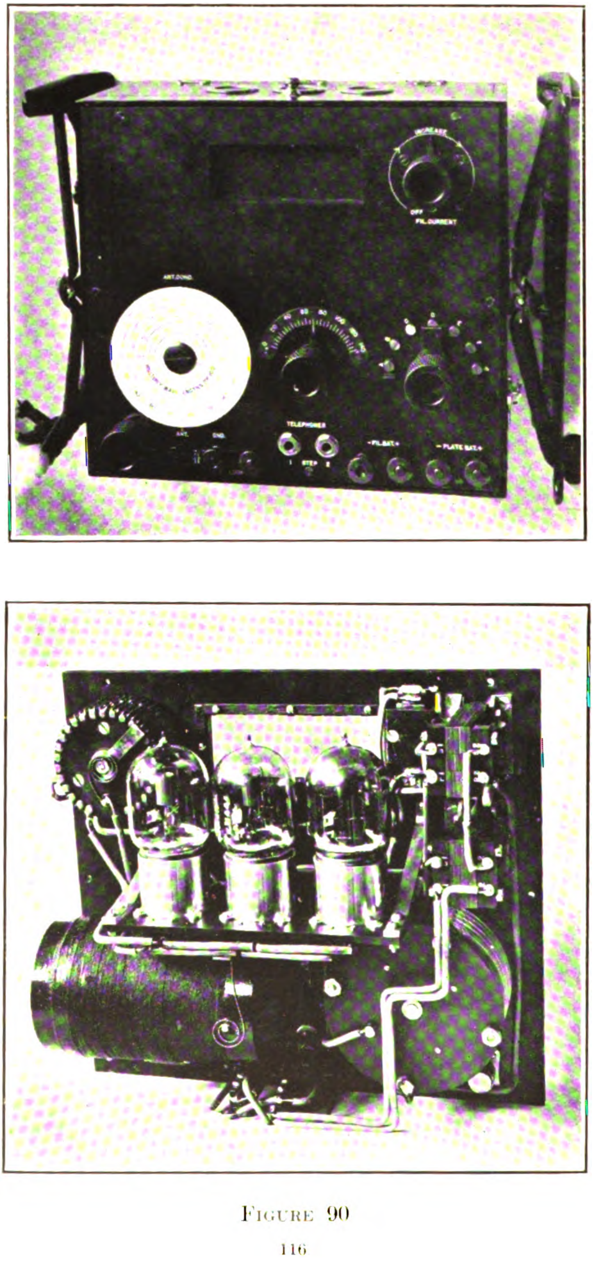

FIGURE 90

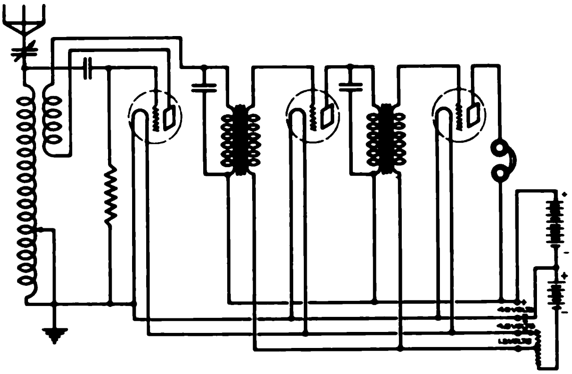

FIGURE 91

ant and because of the close coupling are quite effective over an extremely wide frequency range. In this amplifier a wave length range of 600 to 1,900 meters is obtained. The amplifier is also designed for a wave length range of 1,500 to 5,400 meters, in which case it is known as type SE 1405-B. The schematic circuit diagram of this amplifier is illustrated in Figure 94. Six tubes, type SE 1444, are employed. These tubes are not pro-

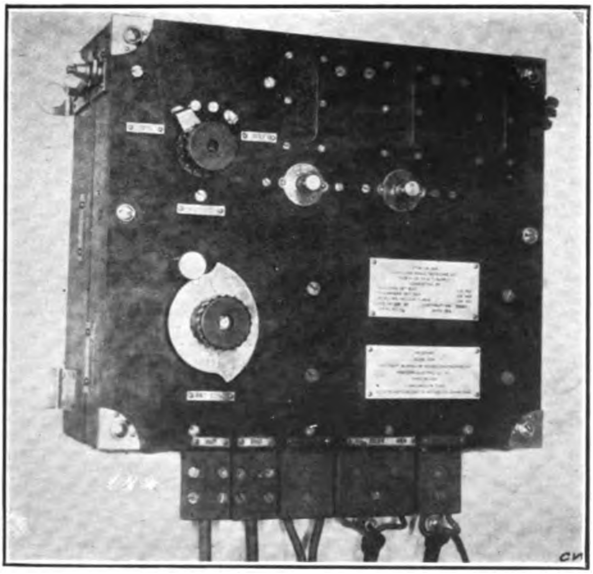

FIGURE 92

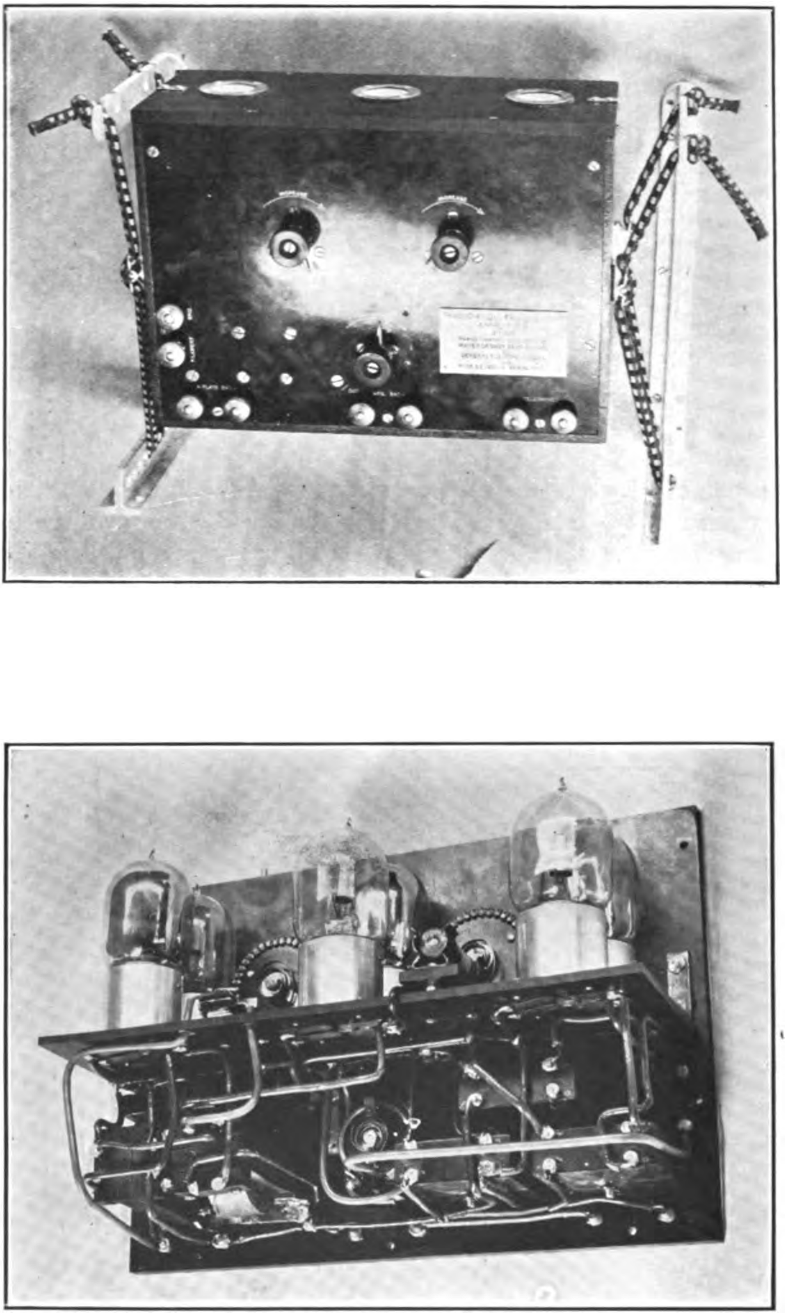

Figure 93

vided with an individual flexible mounting, as the entire amplifier is mounted on an elastic suspension as illustrated. The small inductance provided in the grid lead of the first tube is designed as a choke for undesirable frequencies rising from engine ignition systems.

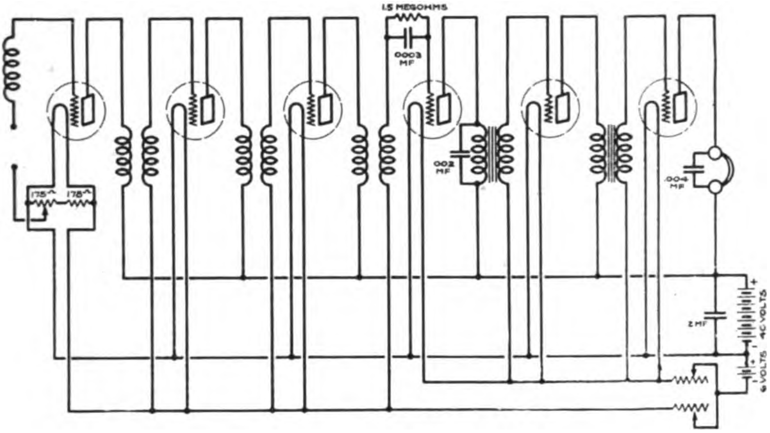

FIGURE 94

The difficulties of quantity manufacture of this amplifier are greatly increased by the fact that the tuning capacities are those of the tubes, and associated apparatus leads, and are therefore difficult to duplicate. This amplifier was designed at the Radio Test Shop, Navy Yard, Washington, D. C., and has been manufactured by the General Electric Company.

In connection with the reception of radio signals on aircraft, interest naturally arises in regard to radio compass or direction finder work. This is one of the most fascinating developments in this field of work, and is one of the most important applications of radio to aircraft.

The method used on naval aircraft is a balanced maximum method. Referring to Figure 95, if "*ABC*" represent a curve of strength of signal as a coil is rotated thru 180 degrees in a field of radio signals, it will be seen that the maximum is quite flat and could not be detected by the ear within very wide limits. Now let "*DEF*" represent the curve of strength of signal for a coil 90 degrees from "*ABC*", or a right angle coil. When "*ABC*" coil is at some angle "*OY*," it will have a strength of signal "*YM*," and "*DEF*" coil will have a strength of signal "*YX*."

Then if the signals "*YM*" and "*YX*" are added there will be a resultant strength of signal "*YN*," and if they are opposed there will be a resultant strength of signal "*YP*." In other words, if the "*DEF*" coil is connected in series with the "*ABC*" coil by means of a reversing switch, it can be made to add or oppose its strength of signal to that of the "*ABC*" coil. The reversal of the "*DEF*" coil is equivalent to advancing its angular position 180 degrees and its curve of signal strength becomes "*FE′D′*". Then when the "*ABC*" coil is at an angle "*OY*," the "*DEF*" coil signal strength, "*YZ*," is negative, and opposes that of the "*ABC*" coil. Thus, when the reversing switch is moved back and forth with the "*ABC*" coil at an angle "*OY*," the resultant signal strength alternates in value from "*YP*" to "*YN*," to total difference of "*PN*," which is decidedly noticeable. When the "*ABC*" coil is at the 90 degree position, the "*DEF*" coil has zero strength of signal and no variation in resultant signal strength is obtained by operating the reversing switch. A very sharp point of reference is thus obtained as from Figure 95 it will be noted that as the "*ABC*" coil approaches the 90 degree position there is a large change in value of the "*DEF*" coil signal strength for a small change in angle and this change in signal strength is doubled by the reversing of connection.

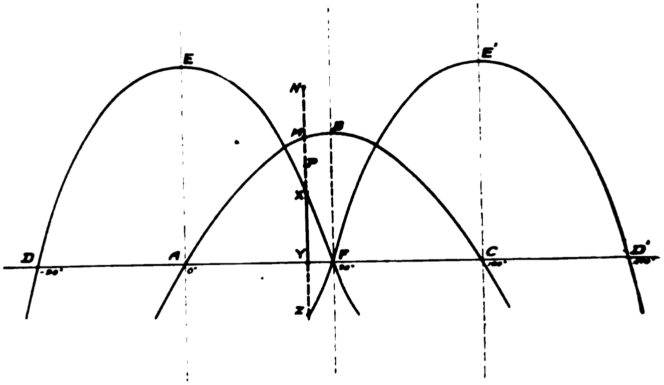

Figure 95

The main advantages of this method over the null method are the facts that there is no danger of mistaking a cessation of signals for the null point, the greater criticalness of this method, and the fact that on account of the use of the higher intensities of signal it is more adaptable for work thru the acoustic disturbances on aircraft.

This type radio compass involves, therefore, a main, or "*A*" coil, an auxiliary, or "*B*" coil, a reversing switch, a compensating device for the auxiliary coil when it is omitted from the circuit, a tuning device for the system, and a receiver for the signals. The compensating device for the auxiliary coil must be provided so that the main coil may be used alone in initially picking up the signal, and so that the entire system may be properly tuned when the auxiliary coil is thrown into circuit for the reversing process.

Aircraft receiver type SE 950, previously described, is provided with the proper switching to convert it into a radio compass receiver. This change is made by throwing the small switch in the upper left portion of the panel shown in Figure 88, and by utilizing the reversing switch in the lower right portion of the panel. An adjustable internal inductance is provided as compensator for the auxiliary coil and is termed an "internal *B* coil." Later in the work, when the amplifier, previously described, was used in connection with the radio compass work, it was found advisable to embody the radio compass controlling devices into the control panel type SE 1441-A, illustrated in Figure 96. In this case a variable capacity is used as the compensating device for the auxiliary coil. The other condenser in the control panel is for tuning the entire coil system. The reversing switch is similar to that provided in the receiver type SE 950, but is of somewhat improved design.

The coils themselves may be arranged either in the wing structures of the seaplanes, involving the swinging of the entire craft in the radio compass work, or the coils may be mounted within the aircraft and be revolvable so that bearings may be taken without changing the general direction in which the aircraft is moving.

In installing fixed coils in the wing structures of aircraft there are used sixteen strands of number 30 B. and S. gauge copper wire* wrapped with one layer of cotton and one layer of para rubber having a total outside diameter of 0.125 inch (0.317 mm.) Several of these wires are laid parallel in linen tape which is glued to the struts and wing surfaces of the flying boats, as illustrated in Figure 97. This form of coil has not been widely used in the Naval air service itself, but has been employed by the Navy in the development of radio compass equipment for air-

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* Diameter of number 30 wire = 0.010 inch = 0.25 mm.

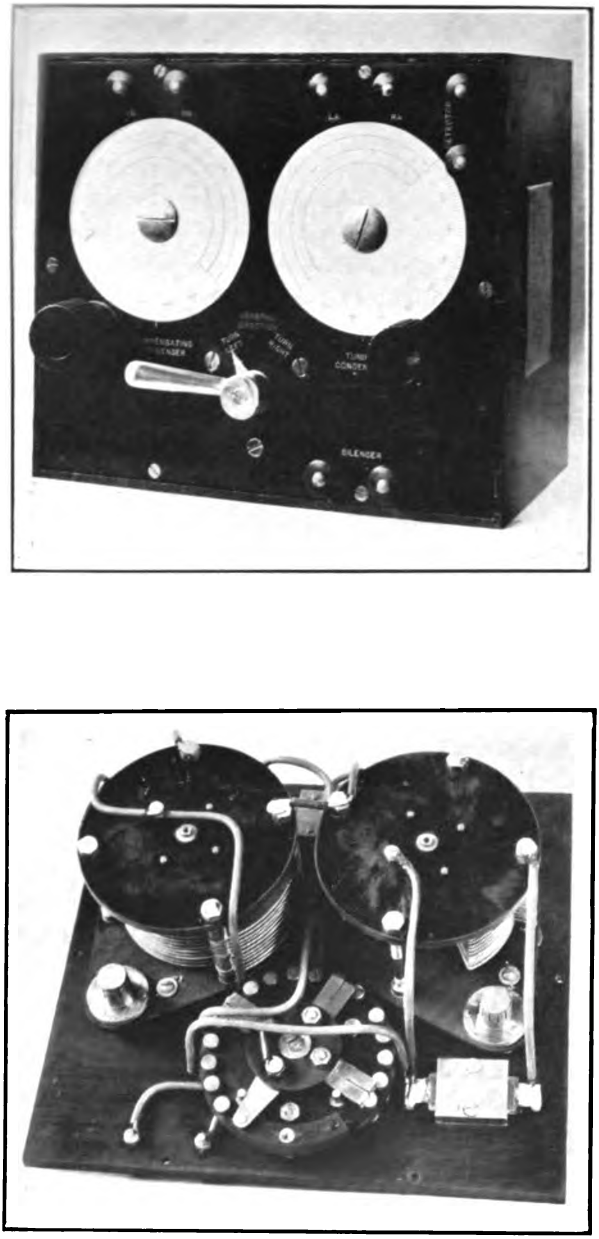

FIGURE 96

planes used by the Post Office Department. In this case satisfactory radio direction finding has been accomplished, utilizing the signals of a 5-kilowatt spark station at a distance of 150 nautical miles (278 kilometers) transmitting on a wave length of 1,500 meters and using radio compass control panel type SE 1441-A in conjunction with an amplifier type SE 1605-B.

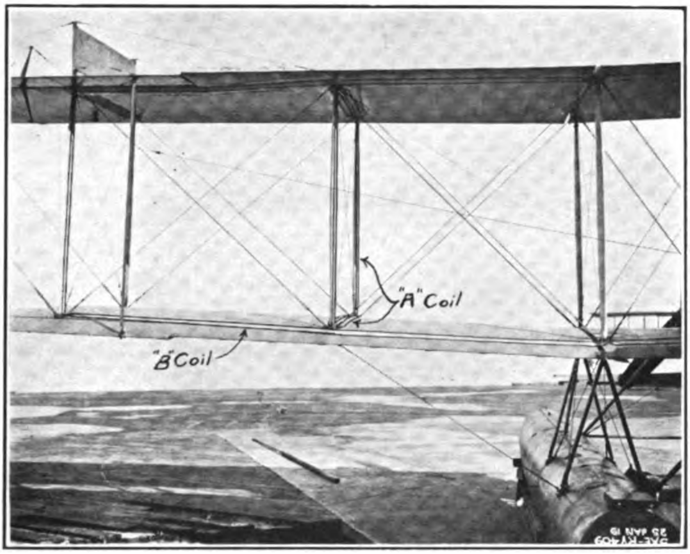

FIGURE 97

Most of the radio compass work on naval aircraft has been performed using revolving coils mounted within the bodies of flying boats and dirigible cars. Experiments have been conducted with revolving coils using the main and auxiliary coil in parallel instead of in series connection. After extended development work it was decided to divide each coil into two sections, connected in parallel, but to use the two coils in series connection.

In the large flying boats these revolving coils are mounted in the tail structure. Figure 98 illustrates coils thus mounted in a type F-5-L flying boat. In this case the coils revolve in the equivalent of a cylinder 34 inches (86.3 cm.) in height and 34 inches (86.3 cm.) in diameter. They are wound with number 18 B. and S. gauge enameled copper wire[*] and are arranged for

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[*] Diameter of number 18 wire = 0.040 inch = 0.102 mm.

[illegible] compass reception on a wave length of 2,500 meters. Used in connection with the standard compass control panel and amplifier, bearings within one degree have been taken at a distance of 300 nautical miles (555 kilometers) on Arlington 100-kilowatt spark signals on 2,500 meters. This is the maximum distance attempted with these signals; the results obtained have indicated that accurate results could be obtained at very much greater distances.

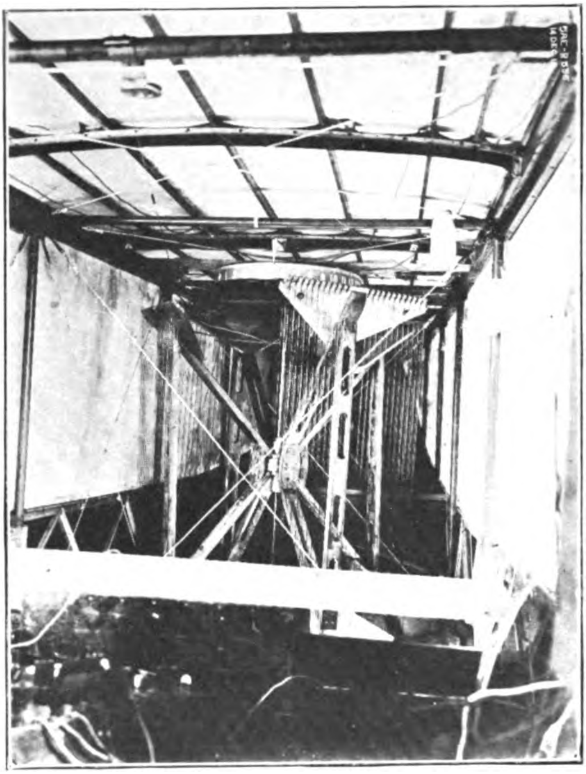

FIGURE 98

Very interesting results have been accomplished using this apparatus with the flying boats resting on the landing runway. In February and March, 1919, five readings within two degrees were taken in this way at Hampton Roads Air Station on spark signals from the 25-kilowatt Swann Island Station, 1,200 nautical miles (2,220 kilometers) distant, on a wave length of 2,428 meters. The useful performance of the aircraft radio compass equipment during the trans-Atlantic flight is described later.

Similar sets of revolving coils are installed as standard equipment on all large flying boats and dirigibles, the sizes varying between the 34 inch (86.3 cm.) size just described, and the size illustrated in Figure 99 which are installed on flying boats type NC and Class "C" dirigibles, and which revolve in the equivalent of a cylinder 40 inches (101.5 cm.) in height and 40 inches (101.5 cm.) in diameter. The International Radio Telegraph Company rendered very valuable assistance in the design of compass coils of the revolving type.

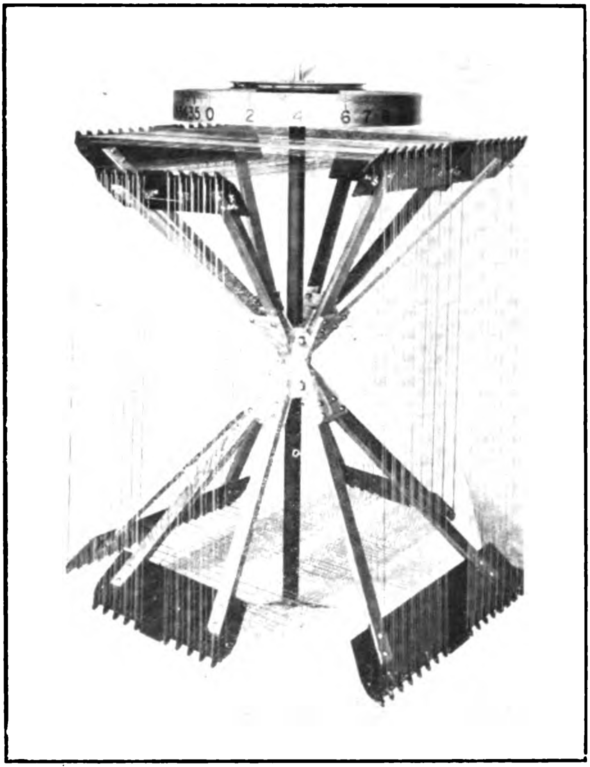

FIGURE 99

With all aircraft radio compass installations there are now

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* Diameter of number 18 wire = 0.040 inch = 0.102 mm.

used the standard radio compass control panel and five step amplifier. All of the coils are constructed in a similar manner, being wound with number 18 B. and S. enameled copper wire*. It has been found that the advisable ratio of turns on the auxiliary coil to those on the main coil is approximately two-to-one. All of the standard aircraft radio compass work is performed on spark signals, it having proved thus far that sufficiently critical bearings cannot be taken on the continuous wave signals.

With the enormously improved radio receiving conditions on aircraft, it is believed that there may be accomplished in the future highly satisfactory results using a minimum method. This would have a decided advantage in that all the inductance of the compass coils would be available in taking maximum readings to determine the approximate location of the transmitting station, whereas with the present balanced maximum method only one third of the total inductance is thus available.

A complete aircraft radio installation which is of particular interest to consider is the special one adapted for the type NC flying boats used on the trans-Atlantic flight. The radio equipment consisted essentially of a 500-watt propeller-driven spark transmitter, type SE 1310, a 5-watt battery-driven telephone transmitter, type SE 1340, an aircraft receiver, type SE 950, modified by removal of the radio compass feature, and a radio compass equipment consisting of a revolving set of coils as illustrated in Figure 99, operating in conjunction with a compass control panel, type SE 1441-A, and an amplifier, type SE 1605-B. Arrangement was made for transmission or reception either on a skid fin antenna, illustrated in Figure 100, or on a single wire trailing antenna.

The radio equipment was located in the after portion of the boat, the compass coils occupying the space equivalent to a cylinder 40 inches (101.5 cm.) in diameter and 40 inches (101.5 cm.) high, completely filling the extreme after compartment. The control wires and electrical leads for the compass coils were brought thru the bulkhead forming this compartment, just forward of which on the starboard side were located the radio instruments as illustrated in Figure 101. To the extreme right in this photograph is located the reel for the trailing antenna. Against the same bulkhead is mounted the variometer assembly for the spark transmitter. A large special send-receive switch is mounted against the side of the boat. Below this on the table is the telephone transmitter. A proper switch was provided by means of which the amplifier might be used in conjunction with the compass equipment or with the standard receiver. The compass was arranged for the most efficient operation on a wave length of 1,500 meters. Operation was also possible on a wave length of 600 meters; by the use of the variable condenser mounted above the amplifier, compass signals could be received on a wave length of 2,500 meters.

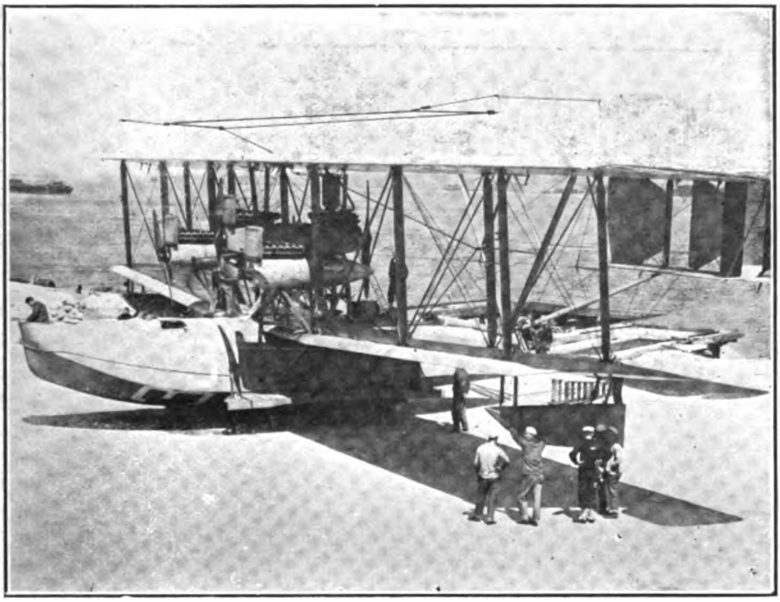

FIGURE 100

The inter-communicating telephone systems on the boats were arranged so that the radio telephone could be used by either the Commanding Officer, located forward in the craft, or the radio operator, thereby permitting radio conversation directly between the Commanding Officers on the various boats. The complete radio equipment installed weighed 200 pounds (90.7 kilograms).

The communication results obtained on the trans-Atlantic flight were most interesting. Transmitting with the 500-watt spark set Type SE 1310, the NC-4 maintained perfect communication with the Cape Race station up to a distance of 650 nautical miles (1,295 km.) and with several destroyers up to 520 nautical miles (1,036 km.). The spark signals from the

NC-4 were copied perfectly by the Naval Radio Station, Otter Cliffs, Bar Harbor, Maine, when the flying boat was 1,400 nautical miles (2,790 km.) distant. The Otter Cliffs station used for receiving an ordinary flat top "T" antenna and a standard vacuum tube receiver with two stages of audio frequency amplification.

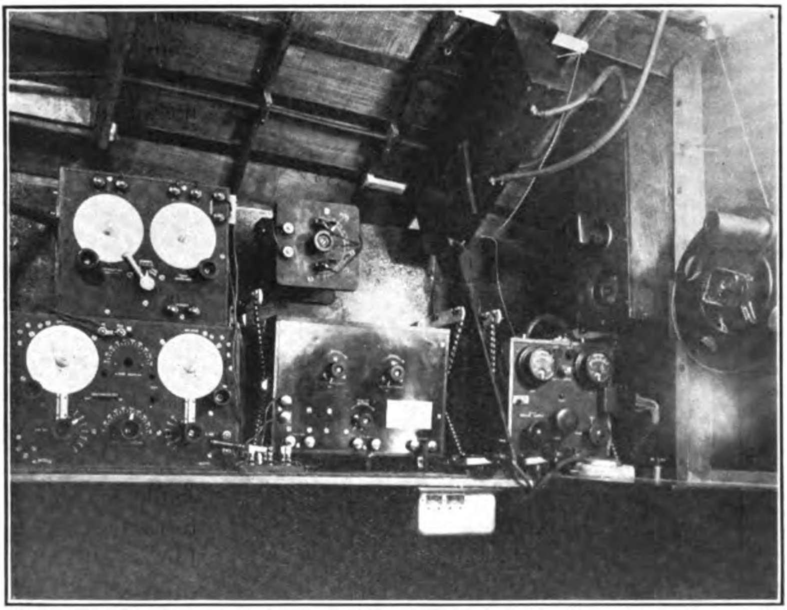

FIGURE 101

The radio compass equipment was used to most distinct advantage when the magnetic compass of the NC-4 became inoperative shortly after leaving Ponta Delgada, Azores, for Lisbon. The NC-4 missed two station ships, at fifty mile (80 km.) intervals, entirely and was badly off course, but radio compass bearings, taken accurately from the signals of the next station ship at a distance of 50 nautical miles (90 km.), brought the flying boat back to the course and made possible the completion of the successful flight to Lisbon.

AIR STATION TELEPHONE TRANSMITTER

For telephone communication with aircraft, naval air stations are to be equipped with a 3.5-kilowatt (input) vacuum tube telephone transmitter type CG 4000, which is illustrated in Figure 102. This transmitter is designed and manufactured by the General Electric Company.

This equipment consists essentially of a panel transmitter, 2.66 feet (0.81 meters) wide, 2.88 feet (0.88 meter) high and 2.17 feet (0.66 meter) deep. Behind this panel six vacuum tubes, type CG 916, are mounted in a spring suspended cradle. Behind the panel are also mounted an antenna inductance, plate and grid circuit inductances, filament rheostats, reactances, switches, and so on. On the front of the panel there are mounted the essential controls consisting of send-receive switch, wave change switch, filament rheostat control, grid coil tuning control, and so on. On the front of the panel are also mounted a voltmeter indicating plate generator voltage, a voltmeter indicating filament voltage, and ammeter indicating plate generator current, an ammeter in the plate circuit of each tube, and an antenna ammeter.

This equipment is arranged for transmission by continuous wave telegraph, telephone, or buzzer-modulated signals on wave lengths of 600, 850, 952, 1,600, and 2,200 meters, utilizing the standard air station antenna the characteristics of which are illustrated in Figure 103. The wave length may be checked by means of a wave meter mounted in the panel, the glow of a lamp indicating resonance.

The schematic diagram for this apparatus is illustrated in Figure 104. Three of the large tubes are used as oscillators and three as modulators. A modulation amplifier utilizing a small 5-watt tube, type CG 1162, is employed. Power is supplied to the equipment from a motor generator set delivering 3.5 kilowatts at 1,500 volts d. c. and current for filament lighting from the alternating current supply thru a transformer.

This transmitter delivers approximately 750 watts to the antenna and will render satisfactory telephone communication from shore to aircraft equipped with standard receiving apparatus up to a distance of 200 nautical miles (371 kilometers). Arrangements have also been made whereby this transmitter in conjunction with receiving apparatus may be remotely controlled and thereby make it possible for the Commanding Officer of an air station, sitting at his own desk, to converse directly with pilots of aircraft operating under his command, and to direct their movements. Such remote control may be conducted over the regular telephone lines by a method of connection shown

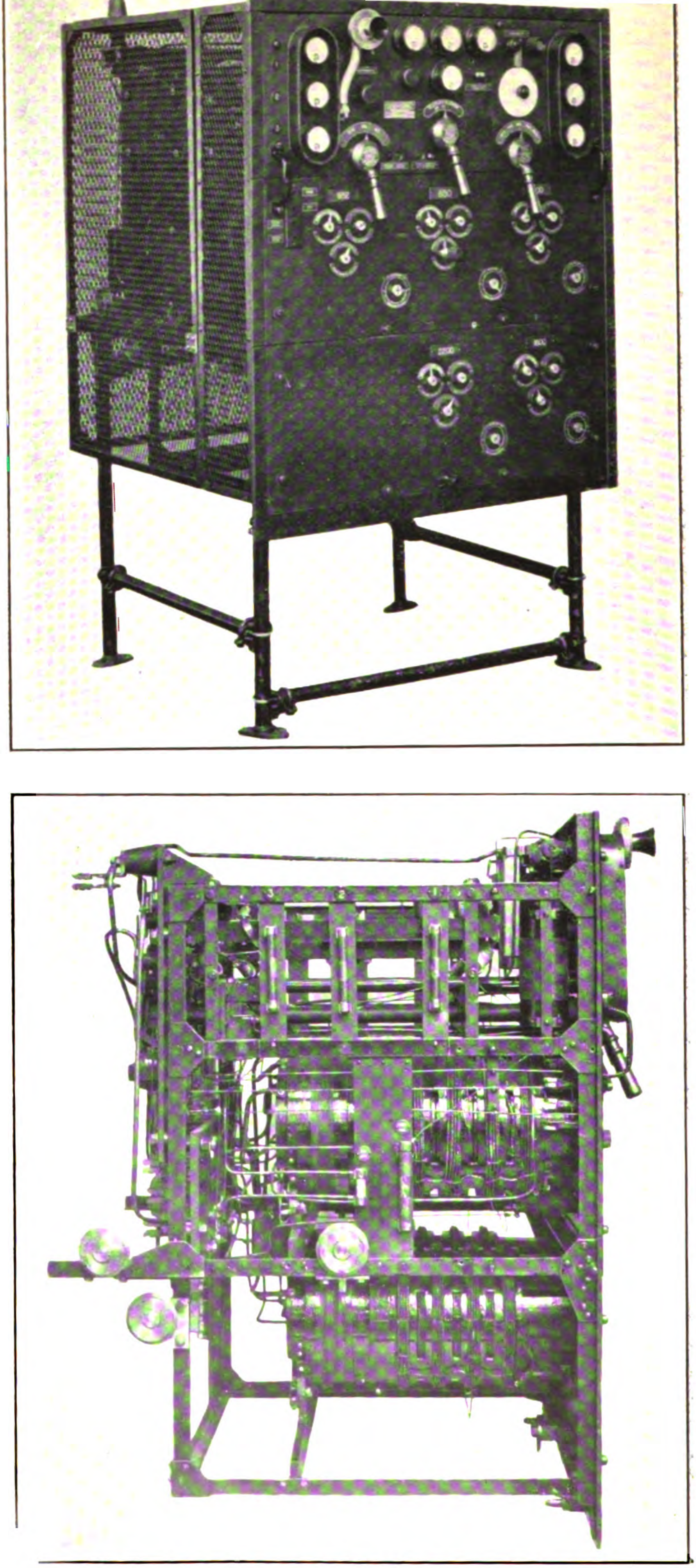

FIGURE 102

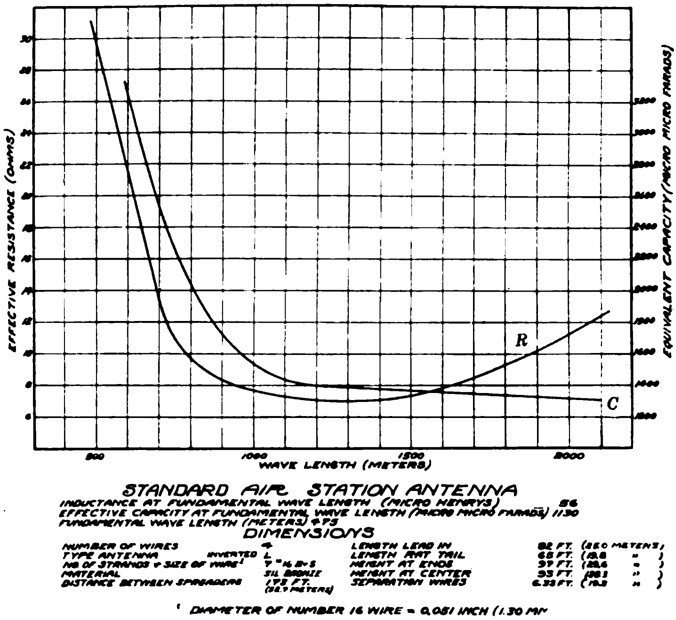

FIGURE 103

in Figure 105. The distant party is placed in communication by being called from the local desk telephone. The remote control of the radio telephone sets may thus be established at any time by closing the connecting switch "A." The receiver

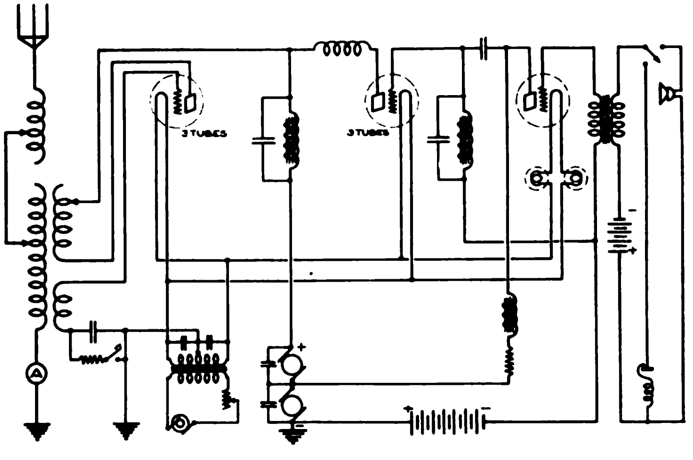

FIGURE 104

on the local telephone may then be replaced on the hook, the opening of switch "A" being sufficient at the close of the remote control conversation to give the local telephone operator the disconnect signal.

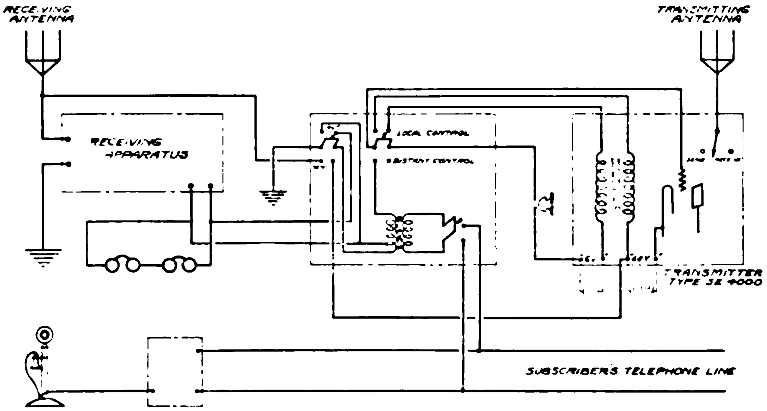

FIGURE 105

A series of tests were conducted during February and March, 1919, with this apparatus in order to decide upon standardized design and investigate the possibilities of remote control. These tests were conducted between the air station telephone transmitter described above, installed at the Navy Yard, Washington, and flying boats operating from the Naval Air Stations, Hampton Roads, Virginia, and Anacostia, District of Columbia. During these tests many conversations were carried on between representatives of the Navy Department seated at their desks in Washington and the radio operators on the flying boats flying over Chesapeake Bay and surrounding waters.

On March 12, 1919, Secretary Daniels of the Navy Department, seated in his office and using his local desk telephone, carried on a conversation with Lieutenant (j. g.) Harry Sadenwater in a type HS-2-L flying boat over Chesapeake Bay at the mouth of the Potomac River, sixty nautical miles (111 kilometers) from Washington. The Secretary's voice was carried over the regular telephone wires to Navy Yard, Washington, where it remotely controlled the output of the air station set located there, transmitting on a wave length of 1,600 meters. The radio telephone signals from the flying boat, on a wave length of 750 meters, were received at the Navy Yard on a Navy standard short wave receiver type SE 1012-A, and after two stages of audio frequency amplification, were carried on the telephone wires to the Secretary's office.

The radio transmitter used on the flying boat was type SE 1380, manufactured by the General Electric Company, and previously described in this paper. An antenna input of 250 watts was obtained on 750 meters. A type SE 950 receiver and a type SE 2000 helmet, both previously described, were utilized in receiving the telephone signals on the flying boat. Figure 106

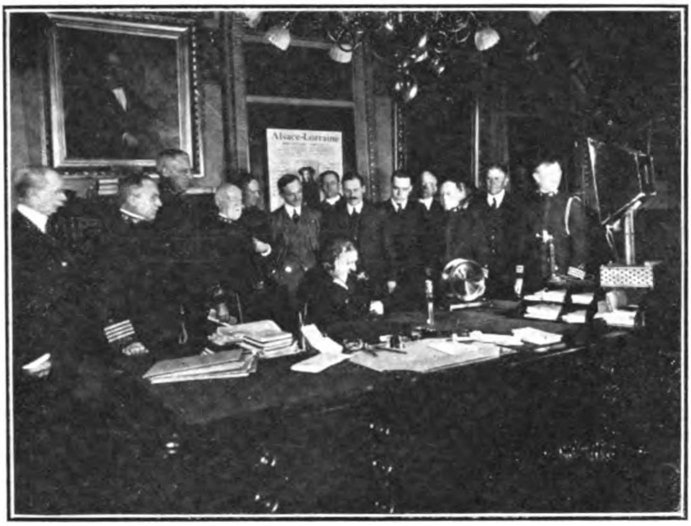

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FIGURE 106

shows Secretary Daniels engaged in conversation, the first of its kind ever conducted over that distance. Figure 107 shows Lieutenant Sadenwater in the flying boat on the return from the trip on which this record was made. Results of the tests conducted at that time indicated the possibility of maintaining perfect remotely controlled conversation with aircraft up to a distance of 150 nautical miles (278 kilometers).

In conclusion, attention is invited to the fact that the progress in naval aircraft radio, involving the development, design, production, and operation of the apparatus described in this paper, has been made within the past two years from a basic knowledge which was extremely meager. The accomplishments

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FIGURE 107

represented in this progress have been due to the individual contributions of the men of the naval service and of the assisting manufacturers. It is indeed difficult to express fitting tribute to these men who have displayed an enthusiasm, an unfailing perseverance, and an *esprit de corps* which reflects the highest credit upon this American contribution to the war activities.

SUMMARY: The application of radio to Naval aircraft is discussed, together with the difficulties of development, standardization, installation, and operation, especially under the stress of war conditions. The first operation of radio on naval aircraft, and the development of the earlier types of sets are described. There is next taken up the development work carried on at the Naval Air Stations at Pensacola, Florida, and Hampton Roads, Virginia.

A brief review of the standard types of naval aircraft is given together with the characteristics of the radiating systems applied to them. After consideration is given to several preliminary matters in connection with transmitting apparatus, the standard transmitting equipments are described, performance data being given.

The standard receiving equipment is then taken up, and the principle, applications, and performance of the naval aircraft radio compass described. In conclusion, consideration is given to the air station telephone transmitter and its use in connection with remote control conversation with aircraft. There is described a recent demonstration of this application for the Secretary of the Navy in which a rather remarkable record for this method of communication was established.

DISCUSSION

John V. L. Hogan: We can hardly say too much in thanking Mr. Johnson for presenting to us a paper of such great interest and value. I wish particularly to express my appreciation of the assistance which the Aircraft Radio Division of the Bureau of Steam Engineering, acting under Mr. Johnson's guidance, has given toward the advance of radio for aeronautic use. From details of design, apparently minor in importance but so often of tremendous consequence in service, right thru to such basic topics as the effective capacity, inductance, and resistance of flying antennas, Mr. Johnson has led us in a most helpful and constructive way. The success of commercial aircraft will depend upon the success of aircraft radio, and so large a contribution to progress as has been given to us by Mr. Johnson deserves the fullest measure of recognition.

Laurens E. Whittemore (by letter): Mr. Johnson has mentioned the fact that the radio direction finder using coils wound upon the struts of the plane has been employed in the development of radio compass equipment for airplanes used by the Post Office Department. In this connection it is interesting to note that in the aerial mail service reliability of service is essential and it frequently becomes necessary to make landings under conditions of poor visibility.

The radio direction finder, as described, can be used by the pilot in finding his way to the general vicinity of the landing field. The exact location of the field cannot be determined by this means since the indications of the direction finder are not dependable when it is above the transmitting station.

It is therefore desirable to have an auxiliary signal which is transmitted from the landing field itself and which is sufficiently localized to enable the observer to determine when he is exactly over the field.

With this object in view, the Bureau of Standards Radio Laboratory has been carrying on experiments at the request of the Post Office Department, and the following are among the results of interest which have been obtained:

A single turn of wire 600 by 800 ft. (196 by 262 m.) was laid upon the ground at the landing field. A 500-cycle current flowing in this wire induces an emf. in a coil of 5 turns of wire wound horizontally around the lower wings of the airplane. The receiving circuit was tuned to 500 cycles, and the signal received using a French 3-stage amplifier. With this arrangement signals were heard at a height of 1,500 feet (500 m.) for a time of seven seconds while the plane was passing over the landing field. Other experiments have been carried on with certain changes in the circuit and in the frequency of the supply current.

A method has also been devised using radio frequency signals. In this case the aerial consists of two horizontal coils 45 ft. (15 m.) square, one placed 20 ft. (6.5 m.) above the other. The current in one coil is in the opposite direction to that in the other. Using a wave length of 1,500 meters, signals have been heard at a height of 2,000 ft. (650 m.) above the landing field.

The electromagnetic field from such a transmitting aerial is in the form of an inverted cone. In this case, while the signals cannot be heard at a point directly above the center of the coils, they can be heard when the plane is in the air and at some distance on either side. Additional experiments along this line are being continued by the Bureau of Standards in co-operation with the Post Office Department.

J. P. Minton: The paper which Mr. Johnson has presented to us is of much importance, and the information given will be most useful in dealing with the problems involved in radio telegraphy and telephony. Mr. Johnson is to be congratulated upon the successful manner in which he has put together the vast amount of information he has collected while engaged in this work for the Navy. The men of the Army and Navy and the engineers and scientists actively co-operating with them deserve the highest commendation for the manner in which they attacked and solved the extremely difficult acoustical problems encountered in airplane telephony. It is in this connection that I wish to speak.

The engineers of the Western Electric Company, co-operating with the Signal Corps, developed during the summer of 1917 the T-1 type of telephone transmitter, which was replaced in the summer of 1918 with an improved type known as the T-3. Since this improved type was not used by the Navy in the comparative tests made by them, it would seem worth while to give briefly some of the characteristics of the T-3 microphone compared with other instruments. The results of some of the comparative tests on various types of airplane transmitters are illustrated in the following table:

Type of Transmitter	Noise Interference (Relative Noise Energy)
T-1	1.0
T-3	0.15
An American Manufacturer's Type	16.
An English Manufacturer's Type	13.
A French Manufacturer's Type	125.

These data are sufficient to show the large range in noise disturbances that is encountered in various types of microphones designed for airplane service. Mr. Johnson has cited the exposed diafram type of transmitter. Comparative tests between this and the T-3 types showed that the instruments were practically equally sensitive to the voice sounds, but the T-3 type was materially less sensitive to the engine and wind noises than was the exposed diafram construction. Subjected to the force of the wind outside the cockpit, the exposed diafram type of transmitter is extremely noisy and at times the effect is sufficiently severe entirely to prevent conversation being carried on. It would seem, therefore, that if comparative tests had been made with the T-3 type of transmitter instead of the T-1, the results obtained and the conclusions reached, as given by Mr. Johnson to-night, would have been materially different.

Considerable has been said regarding the neutralizing effect of the exposed diafram type of transmitter for engine and wind noises. If, however, exposing the diafram on both sides would cause a balancing out of the noises impinging upon the diafram, then it is hard to conceive why this same neutralizing effect is not present outside the cockpit. The noises produced by an airplane during flight are of a most complex nature as regards frequency, phase relations, wave shape, and wave amplitude; and in addition to these factors, the dimensions of the transmitter are comparable with the wave lengths of many of the high frequency disturbances. For these reasons it seems impossible to secure an appreciable neutralizing effect of the kind described in the paper by exposing the diafram on both sides, except for sounds of long wave lengths and these form a minor part of the disturbing noises. Careful tests were made on transmitters, the diaframs of which were exposed in this manner, and it was always found that exposing the diafram on both sides causes greater noise disturbances than when it was protected as in the T-3 type.

With reference to the receiver shields, there are several questions involved. The point upon which Mr. Johnson has laid stress is comfort, and it certainly is a very important part of the problem of developing a satisfactory helmet. The combined receiver shields and helmet described in the paper are essentially like the HS-1 type which was developed and designed by the Western Electric Company in co-operation with the Signal Corps. The helmet, however, described in the paper is inferior to the HS-1 type for telephonic purposes, in that a reduction of the loudness of the speech sounds is caused by placing the receivers at a greater distance from the ears. This, of course, is a serious matter; and, if possible, telephonic efficiency should not be sacrificed for the sake of comfort. That the type of helmet described by Mr. Johnson is not satisfactory from the viewpoint of comfort, when worn over long periods of time, is shown by the fact as reported by Commander Read that the men who flew in the "NC-4" across the Atlantic Ocean found it necessary to discard these helmets for those without telephones because of discomfort.

Our experience has been that these two types of helmets are not satisfactory for sound shielding in large size airplanes equipped with one or more Liberty engines. It was for this reason that the new receiver shields designated "HS-2" were developed by a series of careful tests, the object of which was two-fold. First, by carefully determining the sound absorption, reflection, and transmission characteristics of various materials, we were able to select the most efficient ones for this type of service. Second, by careful tests under actual flying conditions, it was possible so to construct the selected materials that maximum sound shielding, maximum loudness of speech sounds, and maximum comfort were simultaneously obtained. Many tests have been made with the HS-2 type of receiver shields. During the latter half of the summer of 1918, two of the engineers were flying daily as much as four hours a day without any of the discomfort to which Mr. Johnson has referred. It would seem, therefore, that the discomfort to which he has referred could have been eliminated by proper adjustment of the elastic bands and by selecting the proper sizes of helmets to fit over the receiver shields.

T. Johnson, Jr: It has been possible in this paper to give consideration only to apparatus developed to a point at least approaching standardization at the close of the active part of the War, or shortly thereafter. It has been impossible to discuss equipment of very recent development or that which has not been submitted to conclusive tests up to this time.

During the War the only telephone transmitter at all satisfactory for aircraft use in the Navy service was the Magnavox type discussed in this paper. Recent tests of the Western Electric Company's Type T-3 transmitter have indicated that this microphone is of considerable promise, but up to the present time there has not been sufficiently conclusive evidence as regards its performance to merit its standardization by the Navy Department.

As regards the discussion of helmets, examination of Figures 83 and 84 reveal the fact that in the Navy Type helmet the telephone receiver is placed as close, if not more closely, to the ear than in the Western Electric Company's Type HS-1. The same receiver is used in both helmets, but in the Navy Type it lies directly against the ear, altho without undue pressure, while in the Western Electric Company's Type HS-1 helmet a layer of sponge rubber covers the face of the receiver. The Western Electric Company's Type HS-2 helmet has been subjected to thoro test by the Navy Department and has been found unsatisfactory. Regarding the performance of the Navy helmet on the trans-atlantic flight of the "NC-4," it is a fact that Commander Read and Lieutenant Rodd, the radio operator, wore their helmets thruout the entire flight. On this flight the pilots discarded their helmets due to the rather unsatisfactory operation of the intercommunicating telephone system, for it was only in connection with this system that the pilots wore radio helmets. The Navy helmet was worn continuously for thirty-six hours by Lieutenant Esterly, radio operator, on the dirigible "C-5" in its 1,000 mile (1,600 km.) flight to St. Johns, Newfoundland, with entire satisfaction.

B. F. Miessner (by letter): Because of his association with second stage of growth of naval aircraft radio at Pensacola from 1916 to 1918, the writer ventures to record something of the first organized beginnings of aircraft radio in the Navy, as represented by the work done at the Naval Aeronautic Station, Pensacola, under his direction, as Expert Radio Aid for Aviation.

The first problem was that of providing a suitable means for conversational communication between airplane occupants during flight, a problem of great difficulty, as witnessed by the fact that all of the various commercial apparatus and a French Government device, which were submitted for test, had proven unsatisfactory.

Experiments, however, were begun on various schemes for conversational communication, and the problem, when analyzed, resolved itself into two parts, namely, that of developing a suitable sound-proof helmet, and that of developing a sound-transmitting device which would transmit the voice, while eliminating the airplane noises.

The headgear problem was chiefly one of designing suitable forms of airtight cushions for the telephone receivers, and of fitting these into a suitable helmet, so as to provide the maximum of sound-proofness and comfort. While the means at hand did nor permit of great latitude in the selection of materials, a very efficient and fairly comfortable headgear was very soon produced, using both soft rubber eye shields, as used on telescopic gun sights, and commercial forms of telephone ear cushions.

I might add here, that, in my opinion, of all forms of such cushions thus far proposed or used, those employing pneumatic rings of very soft rubber of suitable form, are by far the most comfortable and efficient as sound excluders, as shown by comparative tests of many types.

The anti-noise transmitter problem was of a totally different nature and of much greater difficulty. Renewed efforts were made to devise a transmitter of higher sensitivity and capable of eliminating the noises, while satisfactorily transmitting the voice. These efforts, and the experiments following, terminated in the development of transmitters of both the magneto and the microphone types, which proved exceedingly satisfactory in tests on the highest powered and noisiest seaplanes then available.

These transmitters were named "balanced," or "differential," transmitters in view of the fact that the diafram was exposed equally on both sides to external sound influences.

A great many tests were carried out on this type of transmitter, both in the laboratory and under aircraft noise conditions. At first the most noticeable result of its unique design was to eliminate undesired sounds by making it directionally receptive, so that noises such as motor exhaust, arriving from definite directions, could be eliminated by properly positioning the diafram with relation to the motor.

The directional properties were determined, and voice-noise audibility ratios and articulating qualities were measured with different types and arrangements. Many models of varying forms were tried and mention of the work was made in official reports.

A parallel line of investigation with an acoustic form of communicating set was carried on, altho not so closely, during this period, and this, at about the time of completion of the "balanced," anti-noise telephone work, became so promising in view of its excellent operation and great simplicity, that it was adopted in preference to all electrical devices, and actually manufactured. This device, which is a form of speaking tube, having many special characteristics, was named the "Ductophone," and with some changes in design, has been manufactured and used in large numbers by the Navy Aeronautic Service. This device is being patented, and is now sold commercially under the name "Miessner Airfone."

In view of these facts, therefore, I take exception to Mr. Johnson's statement relative to the origination of this type of anti-noise transmitter by the Magnavox Company of California, and I desire here to state that my interests are fully protected by patent applications and complete data, including dated and witnessed records, establishing very early conception, and other dates.

It may also be of interest here to note, in connection with the statement made by Mr. Minton of the Western Electric Company, the results of some very recent and thoro comparative laboratory tests which have been made with the type No. 340-W-Western Electric Airplane Transmitter, in its original form and as remodelled according to my invention. Under extremely powerful noise influences of the same degree of magnitude as that experienced on high power aircraft, and possessing a very similar noise spectrum, the articulating qualities were approximately 1,000 per cent. better in its remodelled than in its original form.

T. Johnson, Jr. (by letter): The earlier pioneer work conducted by Mr. Miessner at Pensacola was most commendable, but in relation to the equipment actually standardized upon by the Navy for use during the War and subsequently, the work done at Pensacola was of a purely preliminary experimental nature and none of the actual helmets or telephone transmitters developed at that time were ever used in quantities.

THE RADIO TELEGRAPHIC STATION AT ROME (SAN PAOLO)*

By

COMMANDERS B. MICCHIARDI AND G. PESSION

(ROYAL ITALIAN NAVY)

AND

G. VALLAURI

(DIRECTOR OF THE ELECTROTECHNICAL AND RADIO TELEGRAPHIC INSTITUTE OF THE ROYAL ITALIAN NAVY)

1

In the Spring of 1917, the Italian Government decided to carry out, as speedily as possible, the construction of a high-power radio telegraphic station capable of handling heavy and continuous traffic with the Italian colonies on the Red Sea and also to establish radio telegraphic communication with North America. This task was undertaken by the Royal Navy which has, up to the present, been in charge of radio communication between Italy and its colonies, whether of high or low power, or used for commercial or naval traffic.

It was necessary that the new station should have rapid communication with the main offices of the Government, should be situated in readily accessible territory, on moist soil, and in a region which was not either intensely cultivated or covered with trees. In addition, it should be near at least a small town or in the neighborhood of supply stations for electricity and gas. The preceding considerations led to the selection of a tract of land in the Tiber valley in the immediate vicinity of Rome, and specifically, of a stretch of meadow enclosed in a loop or detour formed by the river below the town, immediately below the Basilica of San Paolo.

The two important directions in which communication was to be established are almost opposite, the true bearing along a great circle thru Rome being, for Massaua, 130.5°, and, for New York, 300.9°. The respective distances in the two cases are

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*Received by the Editor, August 25, 1919. Translated from the Italian by the Editor.

3,900 and 6,900 kilometers (2,440 and 4,300 miles). The necessity for establishing this station without delay rendered it impossible to construct new station buildings and it was decided to utilize instead certain existing buildings, namely, those of an old dairy, and to adapt these in the best way to their new use.

The same necessity for haste in establishing the stations was of prime importance in dictating the choice of the system to be used. The first question under consideration, namely, whether the station was to work with damped waves or continuous waves, was solved in favor of the latter, not only because of the intrinsic advantages of that system (which have become increasingly well known in recent years), but also because of rapidity of construction. For similar considerations, the Poulsen arc was chosen without hesitation from among the various generators of continuous waves. The reason for this choice was that the Royal Navy had used the arc with complete success in other small stations, that the arc was very widely used all over the world, and that a device of such simple construction could positively be made ready for use in a much shorter time than any other continuous wave generator. As regards the antenna, account was taken of the well-known necessity for obtaining the greatest possible height in working over long distances, of the convenience of limiting the number of component antennas, and of using an antenna which was not directional in establishing communication with stations situated in widely different directions. On the basis of these considerations, there was provisionally chosen an antenna having the shape of an equilateral triangle, supported by three cables (spreaders) more than 200 meters (610 feet) in height, with the down leads coming from one of the sides. Later, a second set of down leads was brought from the interior of the elevated triangle. In the construction of the antenna supports, taking into account the urgent necessity for haste and not neglecting the desire for economical construction, wooden masts were chosen having a triangular lattice-work structure. Such masts had been previously employed successfully in America, altho of somewhat less height than those required in this station.

Bearing the above considerations in mind, the plans of the station were rapidly worked out in May, 1917; and the masonry work and digging for the adaptation of the existing buildings and for the foundations of the antenna were at once undertaken. In the meantime, construction work was in progress and the machinery and adjuncts were being installed.

The radio transmitting apparatus and its switchboard were furnished by the Radio Engineer, Mr. Cyril F. Elwell, who took care of its construction in Italy; the generators were furnished by the Marelli Company, and the principal controlling switchboard was supplied by the Magrini Company and provided with "C. G. S." measuring instruments. The selection of all this transmitting apparatus was carried out by the personnel of the Navy, which also participated in the construction work and supplied the station with receiving apparatus. The design of the masts and antenna, and supervision of their erection, were given by Mr. Elwell who had previously carried out similar installations in America. The actual work involved in the erection of the antenna was carried out by the Royal Navy, and by masonry workers of the Military Engineering Corps of the Navy.

At the end of October, 1917, that is, in less than five months, the station was ready. Acceptance tests, carried out by the Electrotechnical and Radio Telegraphic Institute of the Royal Navy, were then begun. Immediately afterwards, the station was put into actual operation.

2

The installation of transmitting apparatus is shown in Figure 1. This was located in an old hay loft, the foundations of which had been reinforced, which had been raised until the height of the floor was above the highest flood level of the Tiber, the partitions of which had been completed, and the roof rebuilt. The transmitting room is divided into two parts, one for the generating machinery and the other for the arc.

The generating machinery consists, at present, of a single set without any reserve, but a second unit will be shortly added. The power supply is not at the station, but is obtained from the Società Anglo-Romana as a 3-phase current at 8,500 volts. It is led into a transformer room, in which there is a small transformer of 30 kilovolt-amperes and 8,400-to-220 volts feeding the lighting lines and another transformer of 500 kilovolt-amperes of 8,400-to-525 volts feeding the principal set and all auxiliary motors. This room is on the ground floor of the building and the conductors from it to the transmitting room consist of underground cable. An overload circuit breaker is inserted in the transformer primary.

The motor generator set (Figure 2) consists of two direct-connected machines, having a rigid coupling and mounted on a common base plate. The normal speed of rotation is 645

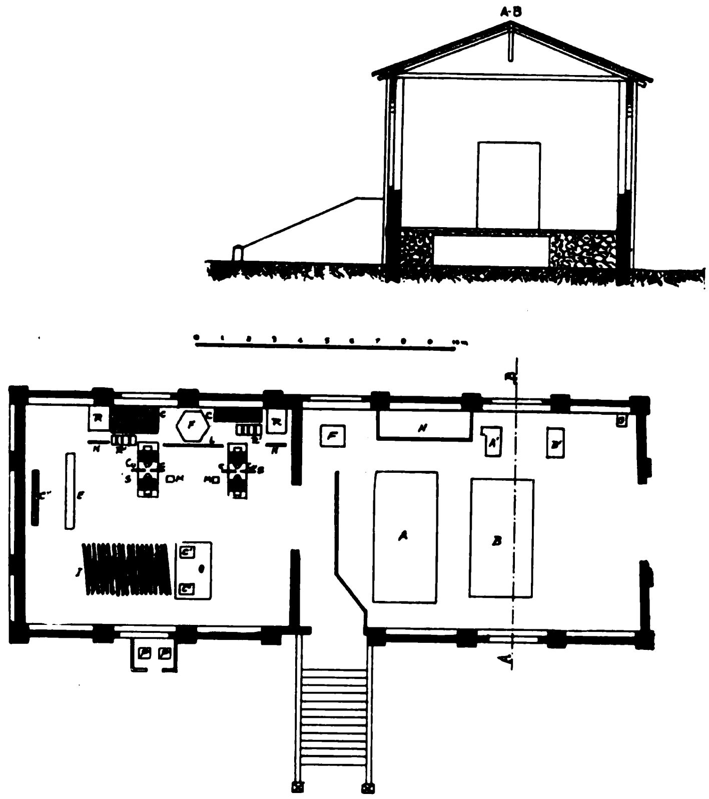

FIGURE 1—Ground Plan and Elevation of Rome Radio Station

A	250 Kw. Set	I	Antenna Inductance
B	200 Kw. Set	L	Arc Switchboard
A′ B′	Auxiliary 110 Volt Set	M	Motor for Rotating Cathodes
C	Shunt Capacity Across Arc	N	Panel Board Carrying Contacts of Starting Arc Resistance
C′	Keys		
C″	Ground Condenser	O	Operating Table Carrying Keys
E	Antenna Switchboard	P	Blowers for Keys
F	Choke Coils	R	Arc Starting Resistances
G	Illuminating Gas Aspirator.	R′	Resistance of Arc Shunt Circuit
H	Motor-Generator Switchboard	S	Poulsen Arcs

revolutions per minute. The motor has the following constants: 387 kilowatts, 500 volts, 530 amperes, 44 cycles, power factor 88 per cent, and full load efficiency 93.5 per cent. The direct current generator has two commutators which can be connected in series or parallel. It is a considerably over-compounded machine, and is separately excited at 110 volts. It was built to meet the following specifications (at full output and with two commutators in series): 350 kilowatts, 1,200 volts, 291 amperes, efficiency 93 per cent. The degree of over-compounding (difference between voltage at full load and voltage at no

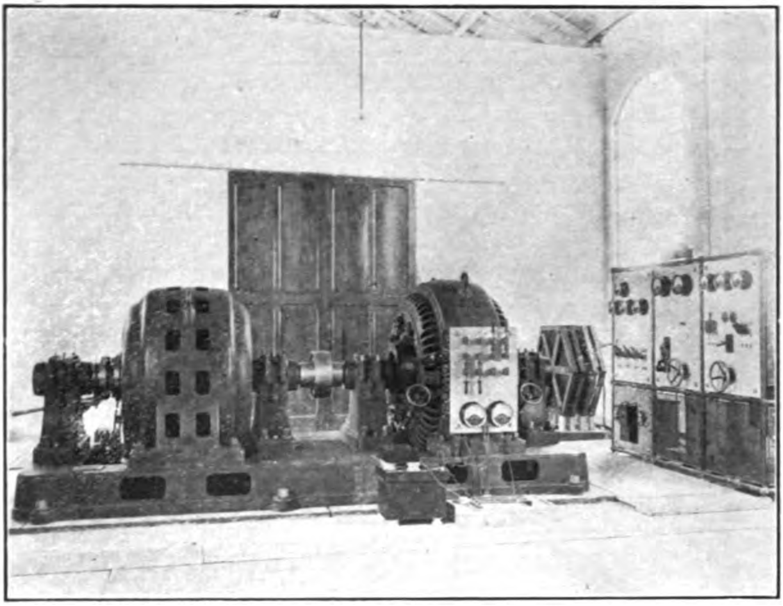

FIGURE 2

load) is naturally dependent upon the greater or less intensity of the separate excitation. In addition, the two halves into which the series excitation winding is divided can be connected in parallel or in series. Thus there can be obtained various characteristics, of which the four pairs shown in Figure 3 constitute examples. The two curves of each pair were obtained respectively with a separate excitation of 2,500 ampere-turns per pole, and with zero separate excitation. The successive pairs of curves refer to the following conditions, starting with the lowest curve and ending with the uppermost:

Commutators in parallel, compound excitation in parallel.
Commutators in parallel, compound excitation in series.
Commutators in series, compound excitation in parallel.
Commutators in series, compound excitation in series.[1]

The windings of the separately excited field will readily give more than 3,500 ampere-turns per pole, and are fed by one or the other of the two auxiliary 110-volt sets. The machinery successfully underwent the acceptance tests carried out according to the standards of the A. E. I., thus showing that the original specifications had been very well met. It was also evident that the machines had been designed with generous dimensions, as was indicated by the fact that the over-heating (beyond ambient temperature) did not even exceed 40° for any electrical part.

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[1] Experiment has demonstrated that it is not necessary nor even convenient to have too high a degree of over-compounding. Indeed, normal operation takes place with the armatures in series and the compound excitation in parallel, this latter being strongly reduced with a shunt of low resistance.

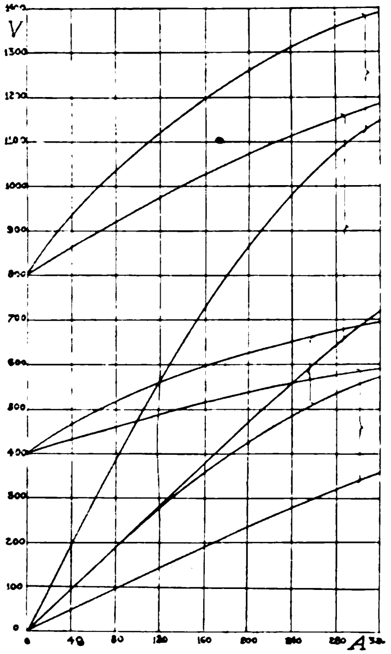

FIGURE 3

In addition to this, taking into account the direct electrical connection between the dynamo and the generator of sustained waves (that is, the arc), a severe test of dielectric strength was carried out on the dynamo windings. This consisted of the application for a 10-minute period of an alternating voltage having the effective value of 5,000 volts.

A simplified schematic drawing of the operating switchboard for controlling the principal unit and the auxiliary units is given in Figure 4. Two over-load circuit breakers were provided, as well as the usual protective apparatus. One of these is of an oil breaker in the 3-phase line, and the other an air breaker in the direct current line. There were also provided every necessary means for protecting the dynamo from radio frequency currents originating in the arc and also against possible grounding of the antenna. Considering further matters of insulation, we find

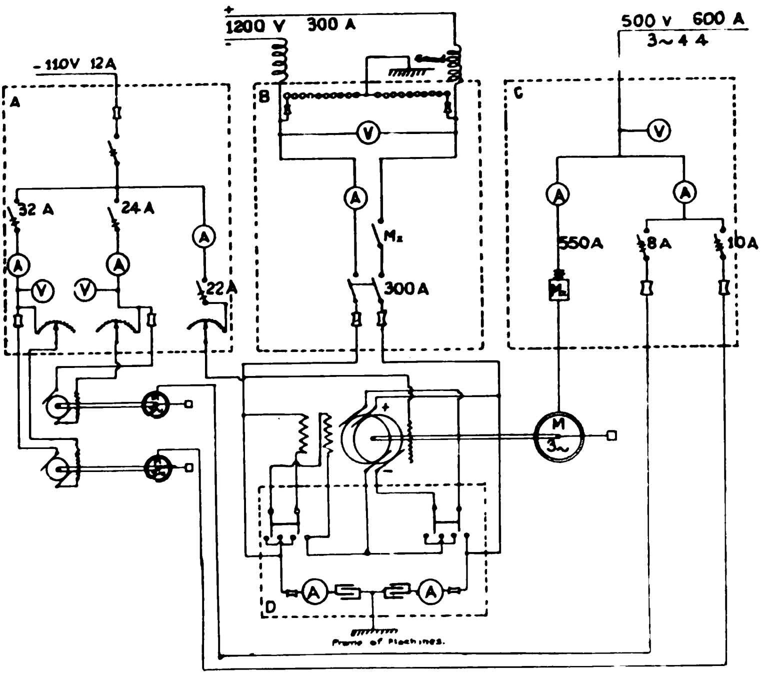

FIGURE 4

that the antenna is separated from ground by a condenser and that it is necessary that the framework of the dynamo, and therefore that of the entire unit, be also insulated from ground. This is accomplished by supporting the base plate of the set on a beech wood framework, the wood having been previously boiled in paraffin in a tank. In this way an insulation resistance between the frame and ground of the order of several tens of megohms is obtained. To prevent any appreciable oscillating current from getting into the windings of the dynamo and thus giving rise to dangerous potential differences therein (1) there are inserted in the supply line of the arc large choke coil inductances and (2) the terminals of this line are connected to each other thru an ohmic resistance (shown in Figure 4 in the upper portion of panel *B*) and consisting of 20 straight filament lamps connected in series with the middle point of the series grounded, and (3) across the terminals of the machine there are shunted a number of condensers divided into two sets, that connected to the positive pole being 0.1 microfarad and that connected to the negative pole being 2 microfarads, and with the junction point of the sets connected to the frames of the motor generator set. This last shunt also includes two ammeters and two fuses, and is properly arranged on the panel *D* (Figure 4) which is mounted on the machine and also carries the series-parallel switches of the armatures and of the series excitation (Figure 2). Under normal operating conditions for the arc, the two ammeters on panels *D* indicate an oscillating current of the order of 0.5 ampere.

3

The assembly of arc equipment (shown in Figures 5 and 6) comprises two similar arcs, an operating switchboard, an antenna inductance, and a table on which are mounted the keys and all other transmitting accessories which are necessary in accordance with the diagram of Figure 7. The two conductors

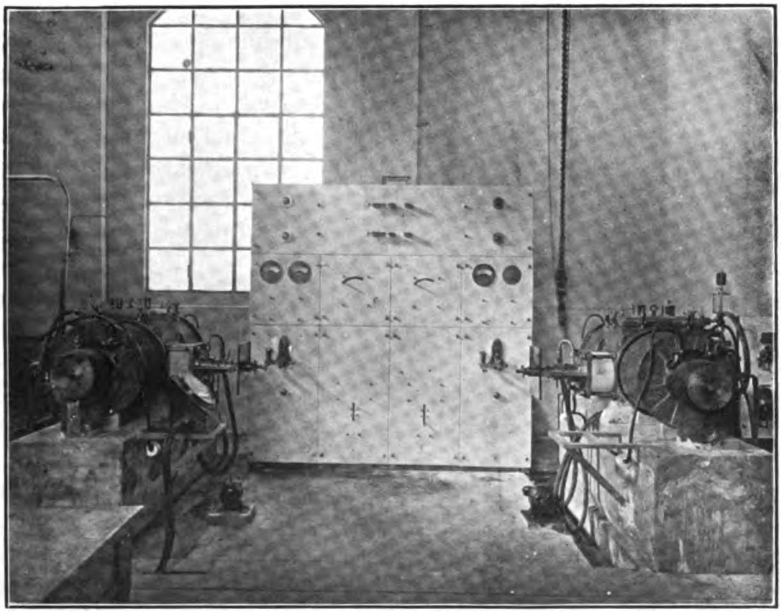

FIGURE 5

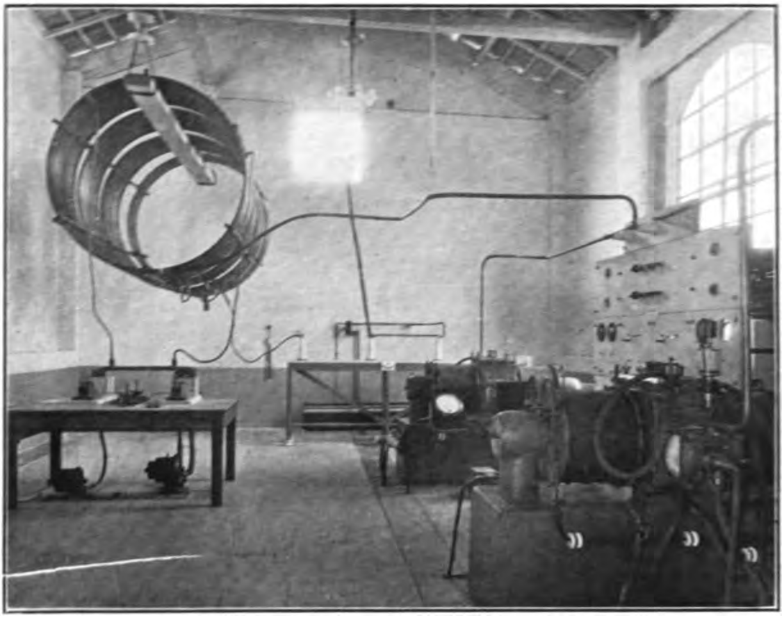

FIGURE 6

of the arc supply circuit (direct current line of 1,200 volts and 300 amperes) before leaving the generator room pass thru two current-limiting inductances, each of which has an inductance of approximately 5 millihenrys (and the reactance of which is still further increased because of mutual inductance).

On the arc switchboards are found a commutator for changing over from one arc to the other, with two extra choke coils and, for each arc, an overload circuit breaker, a starting resistance provided with electric remote control, an ammeter, and a voltmeter. All auxiliary control apparatus is supplied with 110-volt current and connected thru an interlocking arrangement which prevents closing the arc circuit until all necessary preliminaries have been carried out. The diagram shows several of the interlocking circuits.

The Poulsen arc is provided with a rectangular magnetic circuit, the lower portion being enclosed in a masonry base. The magnetizing coils, which are normally used in parallel, are connected to the negative pole of the generator and consist of 356 turns made up of two flat copper strips, each having a cross-section of 0.7 by 0.8 cm. (0.28 by 0.32 inch). The air gap has

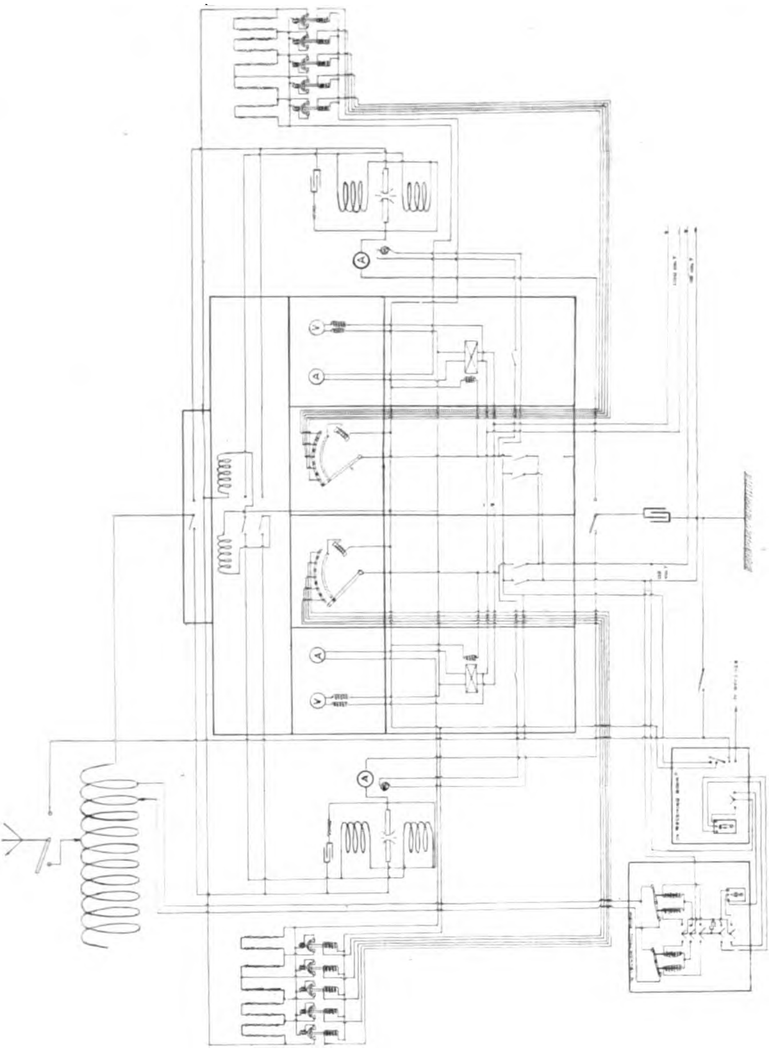

FIGURE 7

a normal length of 8.2 cm. (3.2 inch) and its position relative to the pole pieces and the copper and carbon electrodes is shown in Figure 8.

The intensities of the magnetic field for various exciting currents and different values of air gap were determined. The measurements were carried out, using the ballistic galvanometer method.

The data obtained for the case of an air gap of 8 cm. (3.2

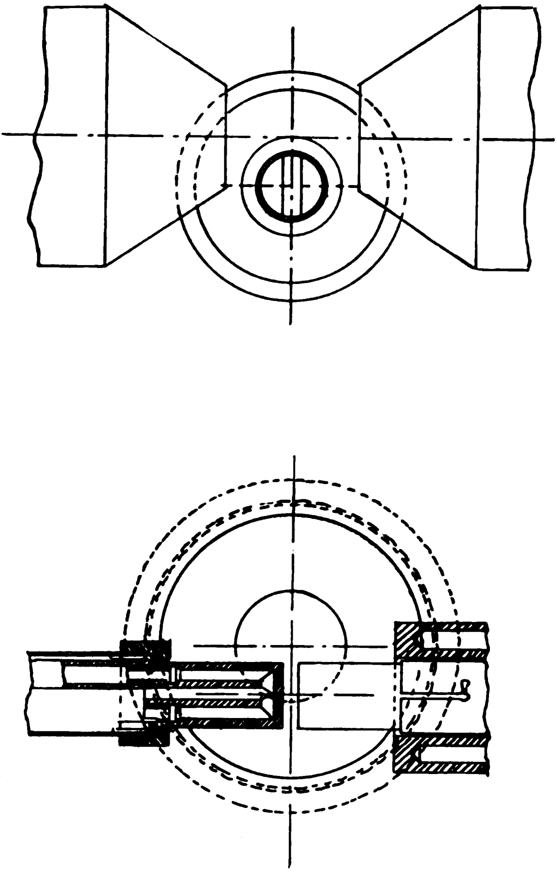

FIGURE 8

inches) is given in Figure 9, which shows the field strength in the region between the pole faces as a function of the distance from the axis passing thru their centers and for various values of the magnetizing current in each coil. Within a range of 15 cm. (6 inches) from this axis, the influence of the yoke is insufficient to distort appreciably the magnetic field, which therefore remains practically symmetrical. In Figure 10, is shown the relation between the field strength H in the center of the gap as a function[2] of the exciting ampere-turns $A\ N$.

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[2] In considering the design of new arcs of higher power, Naval Lieutenant G. Del Santo has carried out a theoretical calculation for the field produced in the arcs at San Paolo on the basis of their geometrical dimensions (Figure 11) and from the supplied magnetization curve of iron. The results of this calculation developed by the method of successive approximations were very satisfactory and proved the possibility of designing such magnetic arc circuits rationally.

Normal operation of this radio telegraphic station is carried on at a wave length of 11,000 meters, with a supply current of 220 amperes, and consequently with a field intensity in the center of the gap of 6,700 gausses.

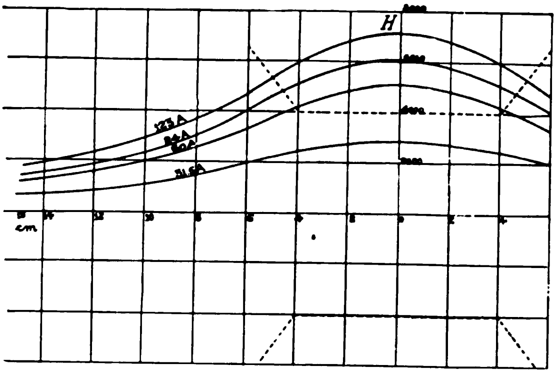

FIGURE 9

When working on a shorter wave of approximately 7,000 meters, the supply current diminishes to 150 amperes and the field intensity to 5,750 gausses.

The copper anode is powerfully cooled and the water, having passed thru this anode, continues on its way to cool the arc chamber.

The carbon cathode is 5 cm. (2 inches) in diameter and is

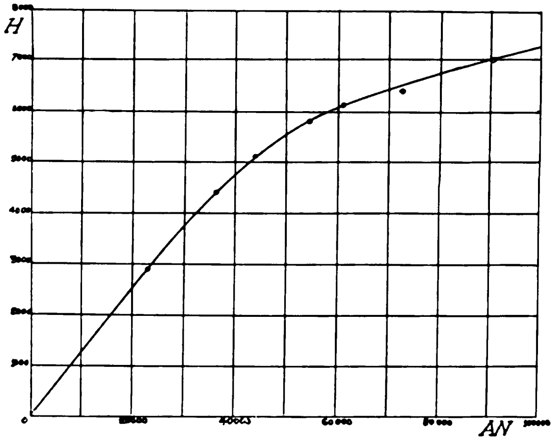

FIGURE 10

slowly turned around on its own axis. Regulation of the distance between the electrodes is carried out by hand.

The arc is struck by bringing the electrodes momentarily into contact; to avoid a short circuit, there is inserted a suitable starting resistance which is gradually cut out as the arc is lengthened. The elimination of this resistance is accomplished by successively short-circuiting its various sections by means of relay-operated contacts which are controlled by appropriate switches on the switchboard in such a way that the electrician can lengthen the arc with one hand and control the starting resistance with the other.

The hydrocarbon atmosphere required for operation of the arc consists of illuminating gas furnished by the Società Anglo-Romana of Rome. The gas enters the arc chamber under a pressure of several centimeters of water and then freely flows thru a tube which carries it outside of the station building.

The anode of the arc is directly connected with the antenna thru a large inductance consisting of 56 turns of copper tubing. The tubing diameter is 3.8 cm. (1.5 inch), and that of the spiral on which it is wound is 2.113 m. (6.44 feet). The length of the conducting tubing is 372 meters (1,135 feet) and its maximum self-inductance 3,060 microhenrys. The cathode is connected thru the antenna ammeter to one terminal of the grounding condenser, the other terminal of which condenser is connected to ground. This condenser consists of an assembly of 80 paper condensers in parallel, each having a capacity of 2 microfarads;

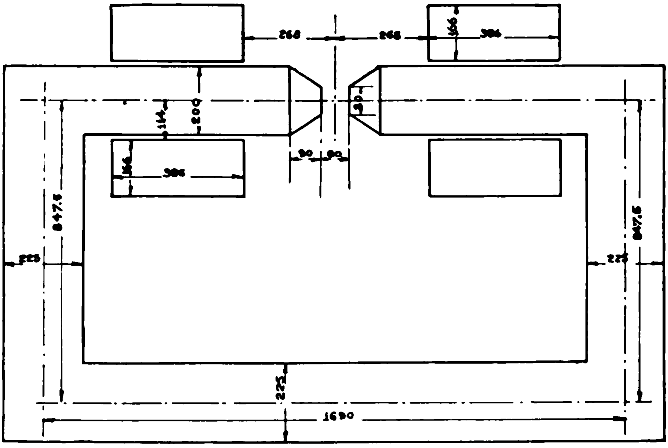

FIGURE 11

and it has the purpose, so far as direct current is concerned, of insulating the antenna from ground, thus avoiding a short circuit of the dynamo in case the antenna should be accidentally grounded.

4

The ground connection consists of 18 plates buried at such a depth as to be in permanently moist earth and of 36 copper wires each 300 meters (985 feet) long, radiating outward from the station.

As has been stated, the antenna is supported on three latticework wooden masts, each 218 meters (714 feet) high, arranged

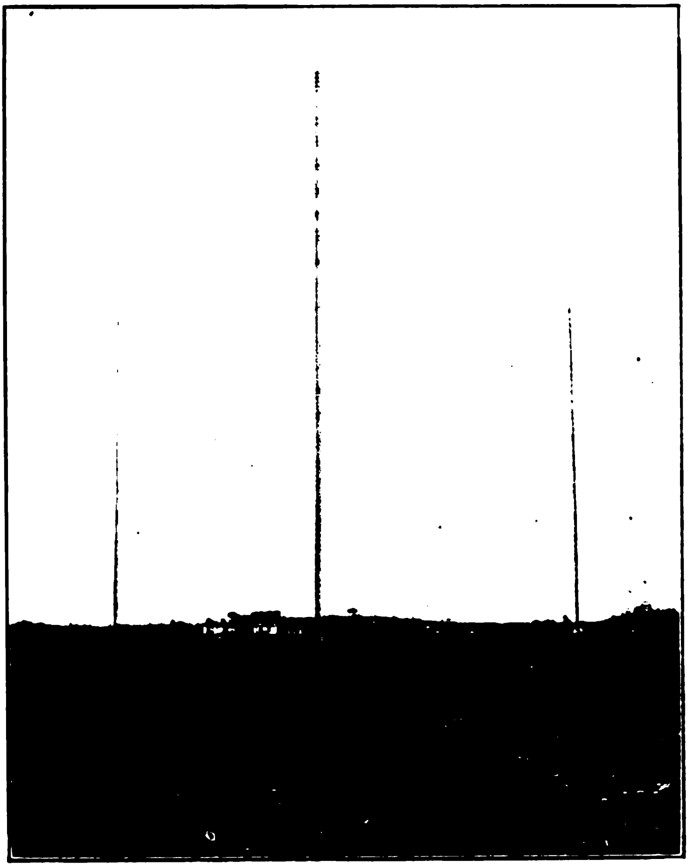

FIGURE 12

at the vertices of an equilateral triangle as shown in Figure 12. The construction of this antenna provided an opportunity to carry on tests and measurements of unusual technical interest, and certain of the results obtained will be given here

It is known that at a given wave length, the current received by a radio station at a definite distance is inversely proportional to the product of the effective height of the transmitting antenna multiplied by the current in its ground lead. By erecting different types of antenna on the same supports, there can be obtained various values of capacity and effective height.[3]

In general, diminished values of effective height correspond to increased values of capacity. Indeed, to increase the capacity the number of antenna wires can be increased within certain limits, which, however, results in increased weight. Consequently, for mechanical reasons, the sag of the wires must be increased, thus diminishing geometrical height and, in general, also the efficiency. A similar result is obtained if the antenna is prolonged by obliquely descending leads or also if the number of wires in the down lead or "rat-tail" of the antenna is increased.

At San Paolo it was, from the first, attempted to obtain a maximum effective height. For this reason an antenna was constructed, made up of a net work of wires stretched between three steel spreaders or cables suspended between the three masts. Considerable tension was placed on the wires to insure that the capacity area of the antenna should be at the greatest possible distance from the ground. Indeed, the tension on the wires was of such value that the catenaries formed by the steel cables connecting the masts were pulled into a practically horizontal plane. The shape of the antenna consequently became that of a sort of triangle with curved sides, these sides being the three steel cables mentioned above (Figure 13). Endeavoring still further to concentrate the antenna capacity area at the greatest height, the down leads were reduced to 13 in number, very close to each other, and were stretched between the radio station and the mid-point of the nearest side of the triangle. The total resulting capacity was 0.0099 microfarads, and it was possible, without trouble, to secure an antenna current of 140 amperes at a wave length of approximately 10,000 meters.

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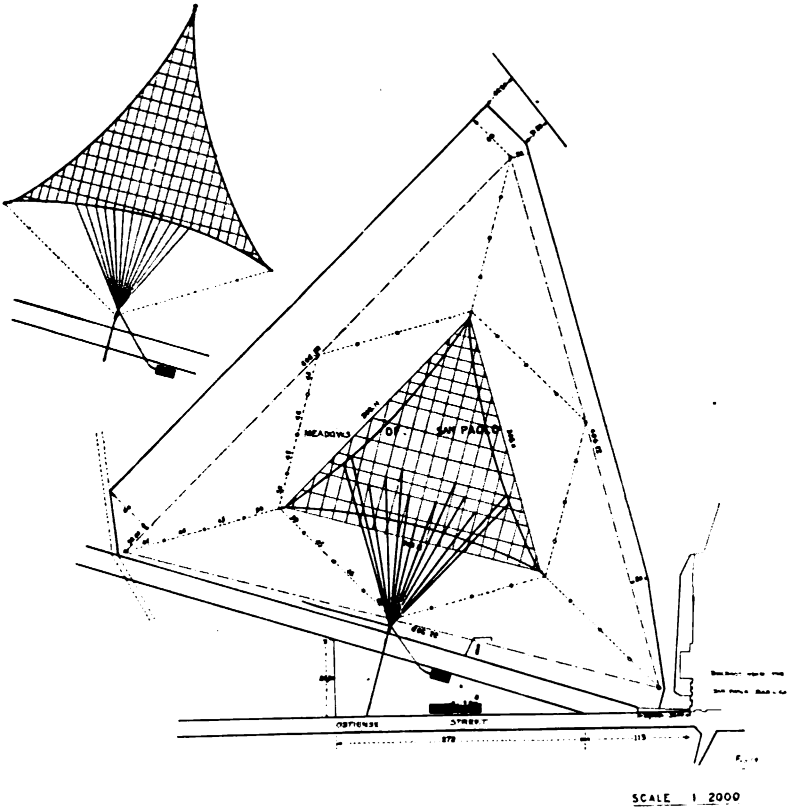
[3] By effective height is meant the height of a straight vertical antenna, the current in which has the same value at all points (usually termed "quasi-stationary distribution") and which produces at a given distance the same electric and magnetic fields as the actual antenna, always provided that the current in the ground lead is the same for both antennas.

Arrangements were then made to measure the effective height of this antenna. As is well known, such measurements really require a determination of the effective value of the oscillating magnetic field produced by the transmitter at a distance from the antenna of several wave lengths.



FIGURES 13–14

To obtain these values, an experimental loop antenna, consisting of a rectangle of copper wire, was set up at Fiumicino in a vertical plane passing thru San Paolo. This latter station was started up and the current induced in the loop antenna, when tuned to resonance, was measured. From the known value of the resistance of this experimental antenna, it was possible to calculate the emf. induced, and therefore, the effective value of the magnetic field. Knowledge of the field intensity permits the calculation of the height of a vertical antenna with quasi-stationary current distribution equivalent to that at San Paolo, or in other words, the effective height of the latter antenna. The value actually found was 151.2 meters (462 feet) for wave lengths between 6,500 and 10,000 meters.

In a second series of trials, the antenna at San Paolo was modified in order to increase its capacity. In contradistinction to the preceding antenna, more wires were stretched across the upper portion of the antenna, thus filling up the empty space between the catenary arcs of the steel supporting cables. In addition, the length of the down lead was considerably increased by connecting to the triangular antenna at various points on its upper surface, instead of connecting to the center of the side of the triangle nearest the stations (Figure 14). Thus the length of the down leads and their distance apart were both increased. The capacity of the new antenna was 0.0112 microfarads and its effective height 138.3 meters (422 feet).

In this second case, to obtain the same voltage as had formerly been secured with the first antenna by 140 amperes, it was necessary to have 158.5 amperes in the antenna. The products of the effective height in kilometers by the current in the antenna in amperes were respectively, in the two cases, 21.2 and 21.9. Thus the second antenna was superior to the first, and was therefore retained. The total resistance of the antenna was measured by the impulse excitation method and therefore with damped current; the results obtained being shown in Figure 15. The actual results which the chosen antenna has given, so far as distance covered is concerned, agree very well with what had

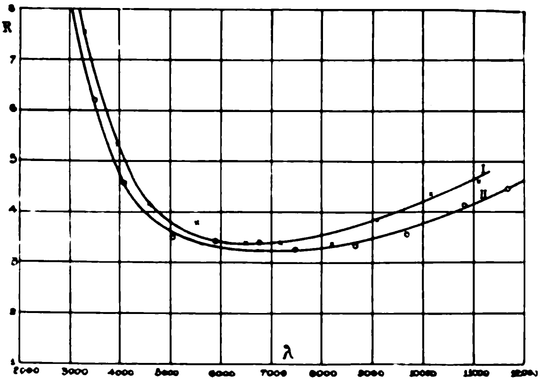


FIGURE 15

been expected, in both dry and rainy weather. The assumption that the resistance measured by the damped current method was probably somewhat larger than the actual value, was confirmed by calorimetric measurements on the arc output,[4] on the basis whereof it was decided that the resistance of the antenna at a wave length of 11,000 meters could not be greater than 3.59 ohms.

5

Signalling is accomplished by short-circuiting a number of turns of the antenna inductance, thus reducing the wave length in the intervals between signals. These two waves, which are called respectively the "working" (sending) wave for the longer and the "stand-by" (compensation) wave for the shorter, differ by 180 meters when the transmitting wave is 11,000 meters. The existence of two waves is a notable disadvantage of the transmitting system adopted since it increases interference with other radio stations. Investigations are in progress on methods of eliminating the compensation wave by the use of an artificial absorbing antenna during the spaces between signals. The short-circuiting of the turns of the spiral is done by a number of relay breaks with large silver contacts cooled by air currents. The relays are controlled by a hand key. In order to use a high-speed interrupter, controlled by a Wheatstone transmitter, experiments are being carried out on pneumatically operated keys. Shunted across the arc when in operation is an oscillatory circuit of short-wave length, consisting of a capacity of the order of 0.005 microfarads and an ohmic resistance which reduces the current in the shunt circuit to approximately two-thirds that in the antenna. The addition of this shunt circuit is a definite advantage in that it permits an increase of about 10 per cent in the effective antenna current at the same supply voltage.

A number of observations were obtained, keeping constant the supply voltage, the arc shunt circuit and the width of the air gap in the magnetic circuit of the arc. In each case the electrodes were adjusted to the best distance apart. A number of quantities were measured as functions of the wave length. In Figure 16 some of the resulting curves are given. For the constant supply voltage of $V_{cc} =$ 800 volts, there are given, the oscillatory current in the antenna I_a, the radiated power P_i, and the "conventional range" D calculated by the Austin for-made on the assumption that a received current of 15 micro-amperes and an effective height of receiving antenna of 100 meters (305 feet) and resistance of 25 ohms are required.

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[4] Further details on these measurements will be given in a later publication.

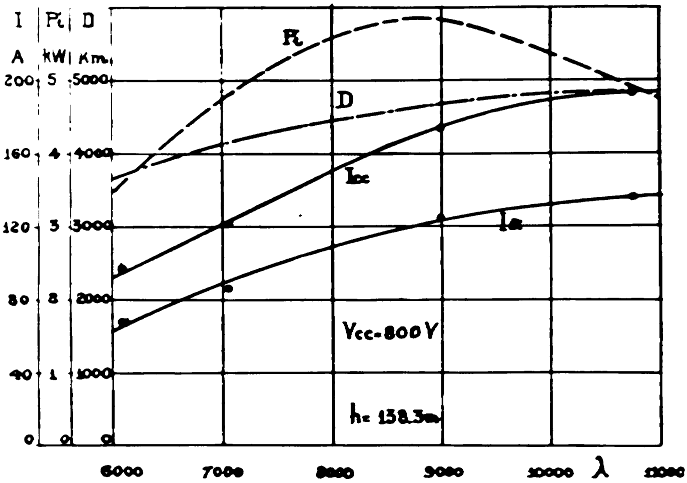

FIGURE 16

6

The radio station at San Paolo has two receiving stations. One of these is of the ordinary type, installed in a small room adjoining the transmitting station. This room is lined with metallic netting connected to ground in order to protect the receiving apparatus against excessive induction from the transmitter. An underground wire joins the receiving room with the transmitting room in order to carry antenna current into the former. There have been installed receivers of the Royal Navy type.[5]

Other underground conductors permit control of the transmitting relays by a key situated in the receiving room.

The second receiving station was intended to permit duplex operation in working with America. This second station is situated near Monte Rotondo, and receives from America by means of a loop antenna, the plane of which points at the American transmitting station and is perpendicular to the line adjoining Monte Rotondo and San Paolo. In consequence, reception is extremely slightly influenced by transmission at San Paolo and thus the American station can be received at any time.[6]

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[5] "L'Elettrotecnica," May 5, 1919; volume 6, page 254.

At Monte Rotondo there are used multi-step three electrode tube amplifiers, some of the French Military Telegraph type, and others of the type of the Royal Italian Navy.

The total number of words transmitted and received in one year from the San Paolo station and at Monte Rotondo (July, 1918 to June, 1919) is 728,249. The cost of this station, in round numbers, is a little less than 1,000,000 Italian lire (200,000 dollars or 40,000 pounds at pre-war exchange rates).

Considering Tuckerton and Sayville, reception from them is not possible after the hours given. The strength of their signals again increases from 9 o'clock up to 1 o'clock in the afternoon.

As regards Annapolis and New Brunswick, altho the signal strength is rather low, it is possible to receive them during these hours. It is not possible to receive them during the afternoon, altho their signals remain, because of the powerful atmospheric disturbances.

In the summer time, the signal strength from all four of these stations becomes low for about an hour before half-past twelve at night, and in winter similarly before eleven o'clock at night. Notwithstanding, reception during these hours is possible.

From June 1st to October, inclusive, reception is difficult from eight o'clock on, because of strong atmospheric disturbances which begin at about three o'clock in the morning and by eight o'clock have reached a considerable intensity which is then still increasing gradually.

November and December are the best months for reception. From January to May there is a gradual increase in atmospheric disturbances which reach the maximum in the summer, as stated above. In any case, however, reception from America was never interrupted, except during times of heavy storm near the receiving station.

Reception is easy enough from Annapolis and New Brunswick, but not from the other two. This may be due to the longer wave length or to the greater signal strength.

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[6] See the tables and information given below for details of this reception.

TABLE 1

VARIATIONS IN RECEPTION FROM JUNE, 1918 TO MAY, 1919 BETWEEN AMERICAN TRANSMITTING STATIONS AND THE RECEIVING STATION AT MONTE ROTONDO.

(Greenwich mean time is used, hours numbered from 0 to 24.) (From midnight to midnight)

Transmitting Station	Wave Length	Positive Reception Begins at Following Time	Months
ANNAPOLIS................	16,900	21	**May, June, July, August**
		20	**September**
		19	**October**
		17	**November**
		18	**January and February**
		20	**March and April**
NEW BRUNSWICK.........	12,800	22	**May, June, July, August**
		21	**September, October**
		19	**November, December**
		20	**January, February**
		21	**March, April**
TUCKERTON...............	9,200	24	May, June, July, August
		23	From September to April, inclusive
SAYVILLE..................	9,600		Same as Tuckerton

The intensity of signals of the above mentioned stations decreases gradually between the following hours:

From 5 to 9 —From June to September
From 6 to 9 —From October to November
From 7 to 8:30 —From November to February
From 8:30 to 9 —From March
From 5 to 9 —From April to May

SUMMARY: The station at San Paolo near Rome is described in detail, including its antenna, down-leads, ground, arcs, controls, and receiving apparatus.

The determination of the most desirable antenna construction is considered; and experimental means of measuring effective antenna height are given.

The methods of simplex and duplex operation are discussed. The reception from four high power American stations is described, and detailed reception data are given.

CALCULATION OF ANTENNA CAPACITY*

By

Louis W. Austin

(United States Naval Radio Research Laboratory, Washington)

The theoretical pre-determination of antenna capacity has always, until recently, been a matter of great uncertainty. Professor Howe[1] in 1914 published an article in the "London Electrician" giving methods for the calculation of capacities of elongated flat top antennas composed of parallel wires, and giving tables and curves making it possible to determine easily the capacities of small antennas of this type. In another article, he later also gave formulas[2] for umbrella antennas.

In 1915, Louis Cohen published antenna capacity formulas in the "Electrical World[3]."

Very recently in Circular number 74, the Bureau of Standards[4] gave formulas for elongated parallel wire antenna capacities which are in very fair agreement with observed values and also with the results of Professor Howe. In all of these formulas, except where the results are given in curves and tables, the calculations are more or less laborious, and, of course, do not apply to antennas of other shapes than those mentioned.

It was discovered empirically during the past year that the capacity of all antennas not too elongated in shape and having their wires not too widely spaced, can be very approximately represented by the formula

$$C=\left(4\sqrt{a}+0.885\frac{a}{h}\right)\cdot 10^{-5} \tag{1}$$

where C is the capacity, a the area, and h the mean height in microfarads and meters. For antennas having a length l more

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* Received by the Editor, July 7, 1919.

[1] G. W. O. Howe, "London Electrician," volume 73, pages 829, 859, 906; 1914.

[2] "London Electrician," volume 75, page 870, 1915.

[3] L. Cohen, "Electrical World," volume 65, page 286, 1915.

[4] Circular Bureau of Standards," number 74, page 239, 1918.

than eight times the breadth b, the above formula must be multiplied by an elongation factor, and we have

$$C=\left(4\sqrt{a}+0.885\frac{a}{h}\right)\left(1+0.015\frac{1}{b}\right)\cdot 10^{-5}$$

Equation (1), while derived empirically, is in reality the sum of the usual expressions for the capacity of a disk in space and that for a two plate condenser, disregarding edge effect. These equations can be depended upon to give results correct in general to 10 per cent for the antenna top, to which must be added the capacity of the down leads and that due to metal towers, and so on. The poorest agreement is found in the case of umbrella antennas, where the amount of wire is often not sufficient to give full capacity.

The closeness of wire spacing required to give approximately full capacity, differs very much with antennas of different shapes and sizes, the required spacing being closer the smaller the antenna. For long parallel wire antennas, this may be calculated from the Bureau of Standards formula already mentioned. In the case of some of the larger antennas, remarkably few wires are required. For example, on a certain triangular antenna about 300 meters (915 ft.) on a side, flat tops composed of five wires about one meter (3.05 ft.) apart, strung around the sides of the triangle gave nearly 90 per cent of the capacity obtained when the whole triangular area was filled in. With parallel wire antennas of medium dimensions, a spacing of one meter (3.05 ft.) will generally insure over 90 per cent of the possible capacity.

Table 1 gives some observed values of capacity for elongated parallel wire antennas, and a comparison of the capacities calculated according to the Bureau of Standards formula and Equation (2). The data for calculation are given below Table 1.

TABLE 1

ELONGATED PARALLEL WIRE ANTENNAS—CAPACITY IN MICROFARADS

Antenna Types	Observed	Leads Est'd	Observed minus leads	Calculated	
				B.S. Formula	Eq. 2
1. Marconi "L"	0.060	0.007	0.053	0.0542	0.0545
2. Medium Ship Type	0.0020	0.0006	0.0014	0.00147	0.00151
3. Portable Ship Type	0.00040	0.00006	0.00034	0.000335	0.000350
4. Model Marconi "L"			0.00028	0.000275	0.000255

DATA FOR CALCULATION

Length m.	Breadth m.	Number of wires n	Approximate wire diameter d mm.	Area m.	Height m.	$\frac{l}{b}$
1. 1830	160	32	10	$29.3.10^4$	104	11.4
2. 91.5	9.15	11	3	$8.38.10^2$	49	10
3. 25.9	1.52	6	3	39.4	12	17
4. 18.3	0.406	6	2.5	7.43	1.52	45

Table 2 gives observed values of capacity for antennas of various shapes compared with the values calculated according to Equation 1.

The chief uncertainty in the observed values of both tables lies in the estimation of the capacity of the down leads, and so on. In the case of the Models 4, 10, and 13, this was measured along with that due to the measuring instruments and subtracted from the observed values.

In addition to its importance in radiotelegraphy, Equation (1) has a more purely scientific interest, since it appears to represent the capacity of plate condensers in general for all values of plate separation provided one of the plates is grounded. The so-called edge effect is represented by the capacity in space term. The expression if exact for circular plates, should be nearly so for all not too elongated forms. Experiments are now being carried on to verify these relations, and the preliminary results indicate the correctness of the equation for all ratios of $\frac{R}{d}$. This part of the work will be published separately in the near future.

TABLE 2

VARIOUS ANTENNA SHAPES—CAPACITY IN MICROFARADS

Antenna Types	Observed	Leads Est'd	Observed minus leads	Calculated Eq. 1	Area m.	Height m.
5. Triangle	0.0155	0.003	0.0125	0.0129	$5.98.10^4$	170
6. Square	0.0180	0.003	0.0150	0.0136	$6.25.10^4$	160
7. Irregular	0.038	0.002	0.036	0.0378	$26.8.10^4$	137
8. Triangle	0.0065	0.001	0.0055	0.0054	$1.156.10^4$	91.5
9. Portable Triangle	0.00084	0.00006	0.00078	0.00085	$2.67.10^2$	12
10. Model Triangle	———	———	0.000075	0.000068	1.457	0.633
11. Model Circular Disk	———	———	0.000135	0.000125	3.56	0.635
12. Umbrella	0.016	0.0015	0.0145	0.0147	$7.24.10^4$	162
13. Model Circular Disk	———	———	0.000053	0.000055	0.933	0.50

SUMMARY: Previously published formulas for the pre-determination of antenna capacity are discussed.

Two new empirical formulas for antennas (one applying to broad antennas, the other to elongated antennas) are given. The results obtained from their use are shown to agree closely with the actual experimentally determined values.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 8 JUNE, 1920 Number 3

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Subscriptions to the PROCEEDINGS are received from non-members at the rate of $1.50 per copy or $9.00 per year. To foreign countries the rates are $1.60 per copy or $9.60 per year. A discount of 25 per cent is allowed to libraries and booksellers. The English distributing agency is "The Electrician Printing and Publishing Company," Fleet Street, London, E. C.

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PUBLISHED BY
THE INSTITUTE OF RADIO ENGINEERS, INC.
THE COLLEGE OF THE CITY OF NEW YORK

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

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THE USE OF GROUND WIRES AT REMOTE CONTROL STATIONS*

By

A. Hoyt Taylor

(Lieutenant-Commander, United States Naval Reserve Force)

and

A. Crossley

(Lieutenant (j. g.), United States Naval Reserve Force)

In previous papers, one of the writers has indicated some of the advantages of a subterranean receiving system for remote control purposes. Those advantages may be briefly enumerated again here.

First. Ground wires have a high degree of directivity which can be utilized for shutting out nearby interference.

Second. For short waves they have an optimum wire length which may be utilized to obtain a far higher degree of selectivity than is possible with an ordinary antenna.

Third. Static is entirely eliminated, permitting successful operation without danger to the operators thru heavy storms.

Fourth. All strays are greatly reduced if it is possible to obtain moist ground for the installation.

Fifth. The ground wires lend themselves readily, under proper conditions of installation, to the use of balanced or stray eliminating systems.

In view of these advantages it seemed possible that a type of remote control might be developed which would not really be remote at all, but could be operated in very close proximity of the sending station. The first experimental work on ground wires with the direct object of studying remote control problems was done at New Orleans by Mr. H. H. Lyon under the supervision of Lieutenant-Commander E. H. Loftin. The first reports of this work were perhaps a little too enthusiastic. Subsequent developments at New Orleans, which will be discussed in detail later, showed that the early investigations at New

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* Received by the Editor, May 24, 1919. Presented before The Institute of Radio Engineers, New York.

Orleans had overlooked one or two important advantages of the ground wire system and also several serious difficulties, which are sure to be encountered in the location of a receiving set within a short distance of a powerful transmitter. The writers, therefore, undertook elaborate experiments at Great Lakes which resulted finally in the building of the Great Lakes remote control station at the foot of the bluff, distant only 600 feet (183 m.) from the nearest tower of the transmitting station. The preliminary work was done in a tent on the beach, the wires being laid north and south as nearly as possible at right angles to a line drawn to the base of the transmitting antennas, of which there were two, both "T" antennas, one suspended from the tops of the two 400-foot (122 m.) towers and the other from points 150 feet (45.8 m.) above the earth. The lower antenna was used on spark work from 600 to 1,500 meters, in connection with a 5-kilowatt Telefunken set, and the larger antenna was used on spark work from 1,500 meters to 3,400 meters and on arc work from 3,000 to 10,000 meters. The arc set was 30 kilowatt and at 6,000 meters would put 50 amperes into the antenna. The wave lengths commonly used were 4,000 meters and 6,000 meters for the arc, and 600 and 1,600 meters for the spark. At these wave lengths, the two antennas, altho one was directly under the other, did not seem to create any serious mutual interference, and it proved to be possible to operate both the spark and arc simultaneously without experiencing any trouble. The desirability of having a remote control station at Great Lakes was brought about by the fact that it was necessary at times to suspend ship-to-shore work on the spark set in order to carry on transmission and reception on long waves. It was deemed desirable to build a remote control station for this purpose, so that spark and arc work might be carried on simultaneously without mutual interference. The preliminary work on the beach developed at once the necessity of having the receiving sets perfectly screened from direct radiation. The tent was therefore covered with wire screening, which had no effect whatever, apparently, until this was made to include the floor and all joints were carefully sewed with wire so as to make a good connection. A very small crack in the door of the screen cage would admit sufficient energy from the nearby transmitter to play havoc with the reception of distant signals. It was even found necessary carefully to screen the leads from the ground wires where they came up into the tent, pieces of screening being wrapped carefully around them to within two inches (5 cm.)

of the receiving set. This screening was connected to the main cage. There seemed to be little difference whether the cage was grounded to the wet sand or whether it was left insulated. Having thus perfected the screening of the receiving sets, both for long wave and short waves, from direct action, the leads of a pair of ground wires were laid as nearly as possible so as to be at right angles to the direction of the transmitter. There was, nevertheless, sufficient asymmetry in the ground to the north and south of the tent, or there was some small deviation from the 90° relationship, so that an undesirable amount of signal, especially on certain wave lengths, penetrated to the receivers. It was found possible to offset this by the use of a very small loop in series with one of the ground wires and placed outside of the tent. By rotating this loop a component in the opposing direction could be picked up, which would nearly neutralize any accidental asymmetry of the ground wires. This loop was only found necessary when transmitting and receiving simultaneously on identical wave lengths. This is a rather difficult feat to accomplish when the nearest transmitting tower is only 600 feet (183 m.) distant. It was, however, very successfully accomplished with the Naval Station at Milwaukee, 50 miles (80 km.) to the north. Signals from Milwaukee were perfectly received and with many times the audibility necessary for readability, at the same time that the Great Lakes transmitter was working ships on 600 meters, that is to say, on the same wave on which Milwaukee transmitted. Simultaneous transmission and reception on identical waves, altho demonstrated by these experiments to be possible, is not recommended as a practical proposition, as the adjustments are admittedly very delicate and the greatest precautions have to be taken to get exact compensation or balancing out of the transmitter signals on the two halves of the underground collecting system. When transmitting and receiving on different wave lengths, altho desirable, it is not absolutely necessary that the receiving wires lie at right angles to the direction of the transmitter. The ground wires have one very remarkable and interesting property when used very close to the transmitter, namely, the electrical impulses which they pick up are not nearly as strong as might be expected. It appears that before the wave obtains the necessary forward slant which makes ground wire reception possible, they must travel some little distance from the transmitter. Perhaps a better way of looking at it would be to say that the ground currents which go with the actual radiation are, in the immediate neighborhood of the antenna, partially offset by true conduction currents. Whatever the reason may be, it is certainly a fact that a ground wire led within a few hundred feet (a hundred meters) of the antenna, does not receive the tremendous signal that one might expect when one compares that with what might be received on an ordinary antenna equally near. For instance, a wire leading straight east into the lake 500 feet (153 m.) long was tuned to 8,600 meters and brought in a signal from New Brunswick of audibility in excess of 10,000. At the same time Great Lakes was transmitting on 6,000 meters with 50 amperes in the antenna, and the audibility of Great Lakes signals was only 200. If it were not for this feature of the ground wires, it would be necessary for the control station, which may perhaps no longer be called remote control station, to be split into two parts, one a short distance south or north of the transmitter, and receiving on east-and-west wires, and the other a short distance east or west of the transmitter and receiving on north-and-south wires. Only in this way could reception from all points of the compass be covered. Thanks to the peculiar lack, or rather relative lack of responsiveness of the ground wires in the immediate vicinity of the transmitter, it is possible to effect a compromise and receive fairly well on a wire which is in line with the transmitter, provided transmission and reception are not attempted on waves which are too close together. During the reception of arc signals on long waves at the beach station in the shielded tent, using north-and-south wires, it was practically impossible to tell when Great Lakes was working on the spark set on either 600 or 1,500 meters. It was to be expected, however, that when an easterly wire was used against the ground, that the spark signals would come in the arc receiving set with tremendous intensity. It was an agreeable surprise to find that the audibility of the Great Lakes spark waves up to 1,500 meters seldom rose to more than 15 on the east-and-west combination. 18 amperes was radiated into the large antenna on the 1,500-meter spark and it always caused greater interference on long waves than on 600-meter spark, but even when using the easterly wires and copying, say Arlington on 6,000 meters or San Diego on 12,500 meters, the operation of the 1,500 meter spark was barely noticeable. An analysis of the Great Lakes traffic showed that most ships were worked either when considerably north of Great Lakes or just after leaving Chicago, in which case they were nearly south of Great Lakes. It was believed therefore that north-and-south wires adjusted for 600 meters would handle all of the necessary ship-to-shore work. Such subsequently proved to be the case. On the other hand, the arc work was all, at that time, east and west, but owing to the non-interference of the spark on the long waves, even when the east wire was used, it was evident that all traffic could be well cared for by locating the remote control station at the foot of the bluff, due east, and 600 feet (183 m.) distant from the nearest tower. Land had been purchased at a point three and a half miles (5.6 km.) distant, but all plans for a station at that place were abandoned, and it was decided by the Bureau of Steam Engineering to put in the new type of remote control, which is no longer remote, at Great Lakes.

REMOTE CONTROL STATION, GREAT LAKES, ILLINOIS

In accordance with recommendations submitted to the Bureau of Steam Engineering, Navy Department, the remote control station was located on the beach, six hundred feet (183 m.) due east from the nearest tower and approximately nine hundred feet (275 m.) from the radio station. Due to the presence of shifting sands, the building was located forty feet (12 m.) back from the edge of the beach, where a hard-pan clay foundation was obtained. To prevent action from the shifting sands and winter ice from disturbing the foundation of the building, a crescent shaped breakwater was built in front of the building and filled in with cinders and rock. To protect the building further and to prevent the face of the bluff from sliding down upon the building from the rear, a retaining wall consisting of wooden piling with 2″×8″ (5.1×20.3 cm.) cross planking was installed ten feet (3.1 m.) back from the remote control station, and the intervening space between the building and the retaining wall was levelled off. A general sketch showing location of the remote control building is given in Figure 1.

The remote control building consists of a one story brick structure 16.5×17.5 feet (5.0×5.3 m.). The interior of the building is divided into four rooms, one operating room 8×10 feet (2.4×3.1 m.), one apparatus room, one clerk's room, and a toilet on the main floor, with a basement containing two rooms, one for fuel storage and the other housing the hot water heating unit. The operating room is electrostatically shielded, the walls, ceiling, and floor of same being covered with metal lath, the doors with number 22 galvanized sheet iron* and the windows with 0.25-inch (6-mm.) mesh, galvanized iron netting. All doors and windows are electrically connected to the walls by means of suitable switches, which complete the electrical screening effect.

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* Thickness of number 22 gauge sheets = 0.025 inch = 0.063 cm.

In order to minimize trouble from high voltage alternating current power leads, one man-hole was built in the bluff 110 feet (33.6 m.) north of the remote control building. The 2,300-volt, three-phase supply mains were led from the local power house by means of underground armored cable into this man-hole, and by means of two transformers, a 110-volt lighting and a 220-volt power source were led into the remote control building by means of two underground armored cables. The power mains supplied current for the battery charging unit which was situated in the basement of the building. All wiring of power and lighting circuits was run in Underwriter's standard conduit.

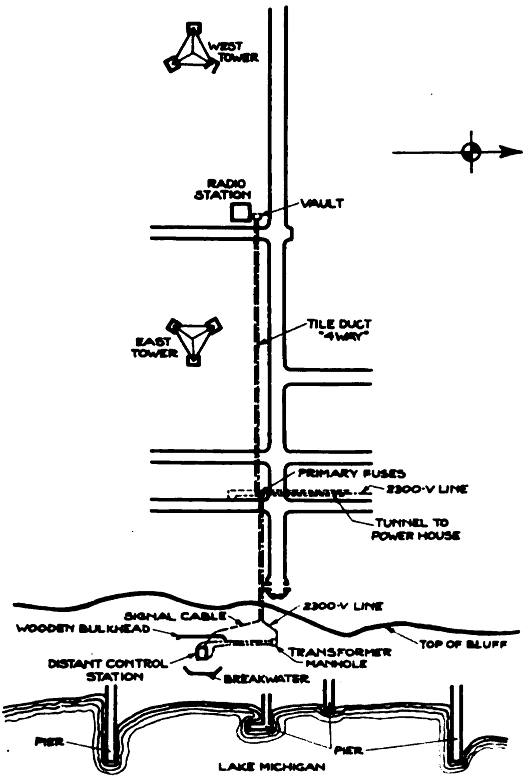

FIGURE 1

To minimize induction further from overhead telegraph lines, all telegraph and telephone leads from the main station to the remote control station were run in a seven-pair standard, paper-insulated, lead-covered, armored, underground telephone cable. The armor of all telegraph, power, and lighting cables was grounded to the water mains upon entering the building. The telephone cable terminates at both the remote control and the main station in an eleven-pair cable terminal box (W. E. Type 14-B). At the remote station, all control and telephone leads from the cable terminal box are run in number 14 duplex, rubber-covered, lead-covered cable* to the various keys and switches. Lead-covered cable was also used at the transmitting station for leads to the various relays and switches. The telegraph equipment for the remote control system consists of six telephone pairs, one pair for each of the following circuits:—spark telegraph, arc telegraph, signal, stop-start switch, local station telephone, and long distance telephone. The spark and arc telegraph circuits control the spark and arc transmitting sets. The signal circuit is used to notify the transmitting station operator when to start and stop the transmitting sets and to change to the different wave lengths. The stop-start switch is connected to the automatic starting panel of the spark transmitting set and allows the remote control operator to operate the spark transmitting set independent of the main station operator. Using 150-ohm relays on various circuits, we found that a voltage of 12 volts was sufficient to operate the relays satisfactorily. Schematic diagram of the control circuits are shown in Figure 2.

The wave lengths used for general work were 600 meters (commercial lake business), 4,000, and 6,000 meters (transcontinental Naval work). Underground and underwater wires of optimum length for these wave lengths were installed. Due to previous experiments on short wave lengths, which proved that the efficiency of reception was increased six-fold when wires were laid in the lake as compared with wires laid in wet sand, it was decided that all wires should be submerged in the lake. This scheme, due to shifting sands, heavy surf, and winter ice, was abandoned for the installation of short wave length wires, with the exception of the due-east 600-meter wire, while for long wave length wires, the substitution of number 2 stranded conductor,† rubber-covered wire for the first one hundred feet (30 m.) for the shore end and of standard underground cable for the rest of the length, proved a very satisfactory installation. The short wave length wires were then installed in the sandy beach at a depth which insured permanent water level. The lake wires were made fast to concrete mushroom anchors, at each two hundred feet (60 m.) distance from shore. These anchors were of the following dimensions—12×12×4 inches (30.5×30.5×10.2 cm.) and were fitted with eye-bolts. The

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* Diameter of number 14 wire = 0.034 inch = 0.162 cm.

† Diameter of number 2 wire = 0.257 inch = 0.655 cm.

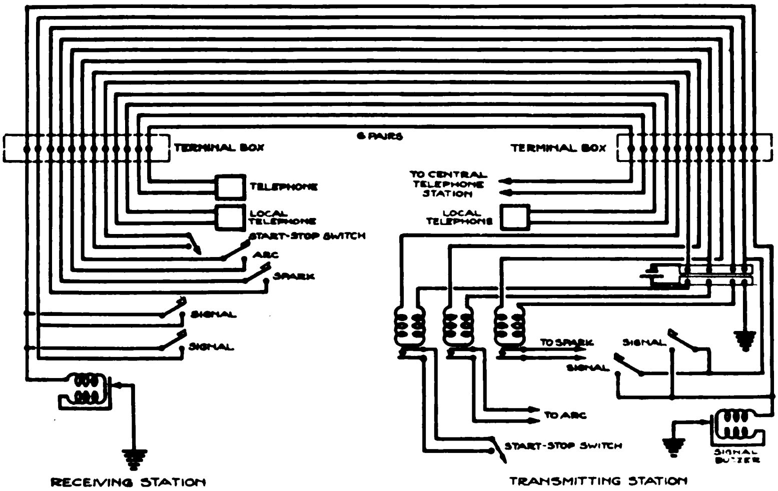

FIGURE 2

underground wire used for the installation consists of seven strands of number 22 tinned copper†, insulated with 0.25-inch (6-mm.) layer of 40 per cent Para rubber and two layers of braid. This wire was given three coats of R. I. W. water-proofing paint, each coat applied evenly by means of two reels and a paint container. The wire was taken from one reel, run thru the paint container, and then wound upon the second reel. Each coat was allowed to dry for three days. See Figure 3 for diagram of container and reels. The above mentioned method of applying paint to wires saves time and trouble and allows the paint to dry without being rubbed off, as is generally the case when paint is applied by ordinary methods. The ends of all underground wires were insulated with rubber tape, and the tape vulcanized and then immersed in a six inch section of

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† Diameter of number 22 wire = 0.025 inch = 0.064 cm.

0.75-inch (1.9-cm.) iron pipe, this pipe being filled with asphaltum. The pipe was then dipped into hot asphaltum, and the pipe and wire for a distance of 12 inches (30.5 cm.) served with marline and the same painted with two coats of R. I. W. paint. The method used in vulcanizing the rubber tape consisted of

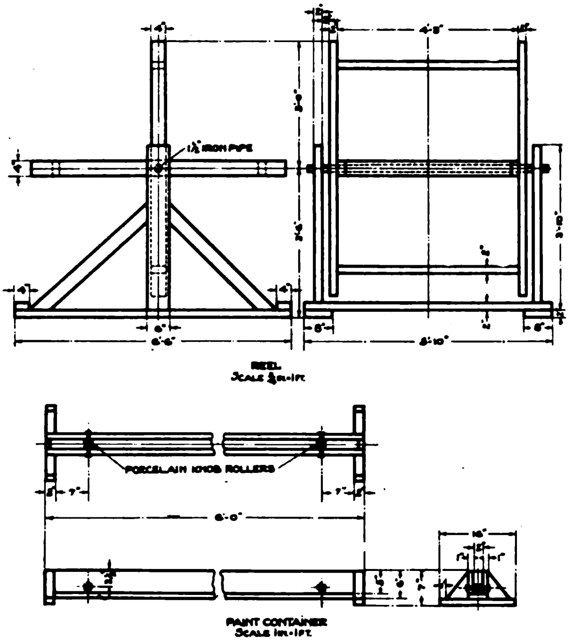

FIGURE 3

holding a lighted paper under the taped joint and revolving the wire, using the hands to compress the heated rubber. The heat from the lighted paper in all cases was sufficient to vulcanize any size joint, and served our purpose very well. To facilitate testing and repairing underground wires, all wires were led to an outlet box situated twenty feet (6 m.) from the water's edge and at the base of the breakwater. Leading from this box to the terminal manhole is a wooden duct, twenty feet (6 m.) long, twelve inches (30.5 cm.) high and twenty inches (50.8 cm.) wide. This duct is approximately nine feet (2.7 m.) underground. The terminal man-hole is made of red brick and is four feet (1.2 m.) square and ten feet (3.1 m.) deep, the top and sides being electrostatically screened and the top fitted with a circular iron man-hole plate. The underground wires were brought into the man-hole, from thence were run thru conduit to the operating tables. The short wave length wires were separated from the long wave length wires and brought into the building at different levels, in order to minimize interference between wires when working, both receiving sets using autodyne reception. The long wave lengths wires terminated at the selector switch on the arc receiving operator's table, while the short wave length wires terminated on the spark receiving operator's table. The selector switch consists of a series of nickel-plated brass studs placed in a semi-circular form and spaced a short distance apart from each other on a bakelite panel. Situated in the center of this semi-circular form are two movable contact fingers, insulated from each other and connected to two separate binding posts at the base of the panel. The different underground wires are soldered to the various studs and by means of the movable contact fingers, any combination of underground wires may be obtained. From the binding posts on the selector switch leads are run to the receiving set. The receiving apparatus consists of the regular inductively coupled receiver, an audion detector and a two stage audio-frequency amplifier. For continuous wave reception a so-called tickler coupling is used between the secondary of the receiver and the plate circuit of the audion detector.

As a "safety first" measure, should an accident occur due to ice floes during the early spring damaging our long wave length lake wires, a rectangle or loop consisting of twelve turns of antenna wire spaced nine inches (22.9 cm.) apart, eighty feet (24.4 m.) long by twenty feet high (6.1 m.), was constructed. This rectangle was situated fifteen feet (4.6 m.) south of the building, with its plane pointing towards Mare Island, California. Leads from this rectangle were led into the building to the selector switch in armored cable ("BX" Type). To prevent bringing external interference into the building from the rectangle when same was not in use, a double-pole double-throw switch was inserted in series with the rectangle leads. By means of this switch the rectangle can be grounded or connected to the receiving apparatus at will. This switch was placed in a weatherproof box on the outer south wall of the building. Figure 4 shows plan of underground system and rectangle.

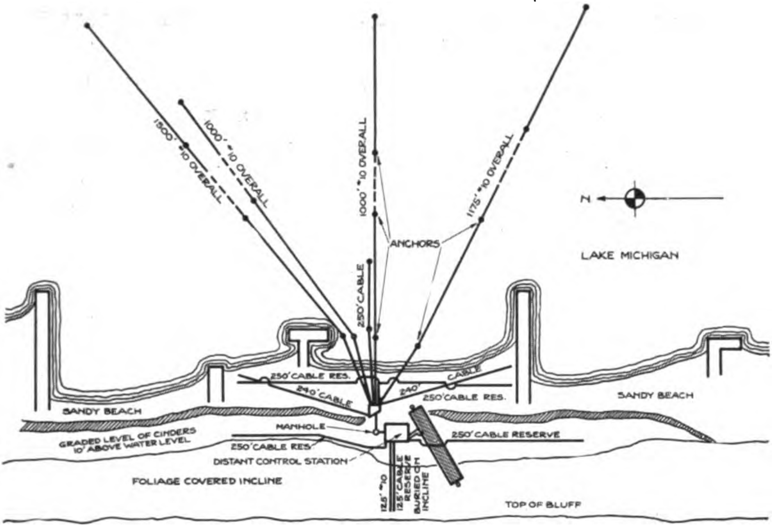

FIGURE 4

REMOTE CONTROL RADIO STATION, NAVAL BASE, HAMPTON ROADS, VIRGINIA

Due to the enormous amount of radio traffic thrust upon the Norfolk Navy Yard radio station and the inability of this station to handle this traffic, the Bureau of Steam Engineering, Navy Department, decided to control remotely the Navy Yard station from the site of the old Jamestown Exposition, a distance of 7.8 miles (12.5 km.) north of the Navy Yard. The underground system having proved efficient for remote control work, it was decided to install the same at the Naval Base to control remotely two 5-kilowatt spark and one 30-kilowatt arc transmitting sets at the Navy Yard.

The site for the control station was selected by the local radio organization, the same being three rooms on the second floor of the administration building and the Bureau of Steam Engineerings' representative had to fit the installation to the alloted quarters. One concrete man-hole of hexagonal shape, six feet (1.8 m.) in diameter was constructed in the basement of the above mentioned building directly under the center receiving room. From this man-hole trenches were dug in the following directions —East, West, North 6° East, and South 40° East, to a depth that insured permanent water level. The water level obtained in the various trenches varied from 3.5 to 9 feet (1.1 to 2.7 m.), and consequently the depth of each trench varied in accordance with these depths.

In order to facilitate quick repair to underground wires, concrete manholes 2×3.5×6 feet (0.6×1.1×1.8 m.) were situated 490 feet (149 m.) apart in each direction. These man-holes were fitted with concrete covers, which, when laid in position, were flush with the graded level of the ground in that vicinity. Figure 5 shows detail of external manholes and center man-hole.

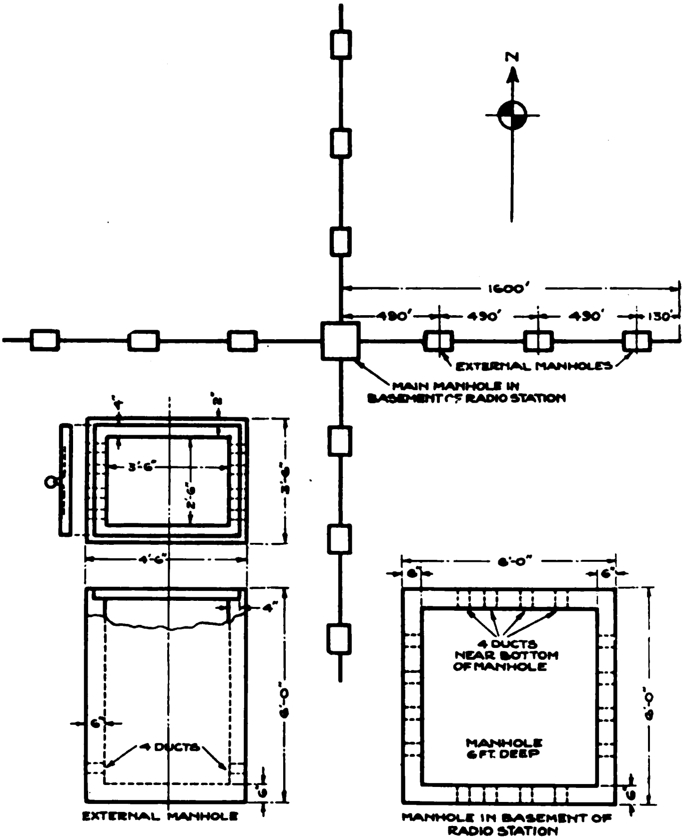

FIGURE 5

The general scheme of the underground wires used consists of five wires buried in each trench, two 600-meter wires 250 feet (76 m.) in length, one 952-meter wire 387 feet (118 m.) in length, one 1,600-meter wire 650 feet (198 m.) in length and one 4,000-meter wire 1,600 feet (488 m.) in length. The underground wires used at the Naval Base are of similar construction to the wires used at Great Lakes, with the exception that the Norfolk wires have an extra insulation of three layers of empire cloth. These extra layers of empire cloth are located between the rubber and braid. The Norfolk wire is classified in the Navy stock list as the "Standard Packard Ignition Cable." All underground wires were painted similarly to the wires at Great Lakes before being buried, and in all cases splices were made in the various man-holes. After all wires were laid and before the trenches were back-filled, tests were made on wires for grounds, using a volt meter and a 110-volt direct current testing source. Any wire showing a ground exceeding one volt was replaced or cause for poor insulation remedied. The general lay-out of the underground wires, together with a plan of the wiring of the main man-hole, is given in Figure 6.

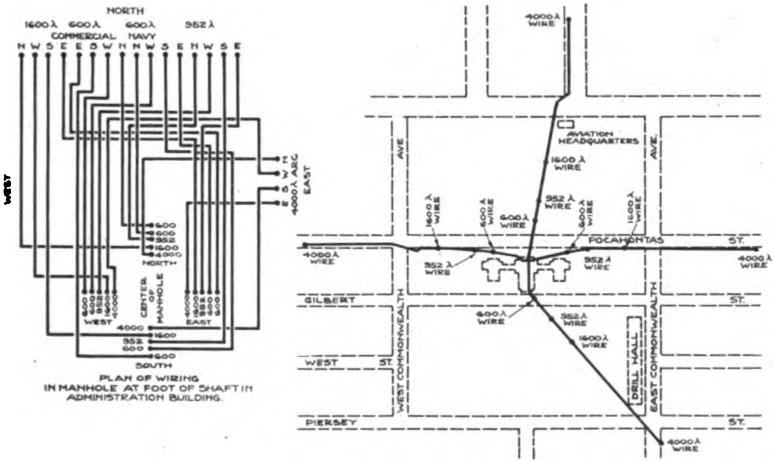

FIGURE 6

From the main man-hole, the wires are led up thru a wooden shaft and under the flooring of the second story to the different receiving rooms. All wires after leaving the manhole are spliced to lengths of number 10 rubber covered wires* and these lengths insulated from the shaft by porcelain knobs, these knobs being secured to the woodwork with brass screws. Where wires cross beams or lead thru flooring, loom is used for insulation purposes and for protection from abrasion between the floor and the operating table, the wires were run thru 0.375-inch (9-mm.) brass tubing. Figure 7 shows general scheme of wiring to the different selector switches in the various receiving rooms at Norfolk.

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* Diameter of number 10 wire = 0.101 inch = 0.258 cm.

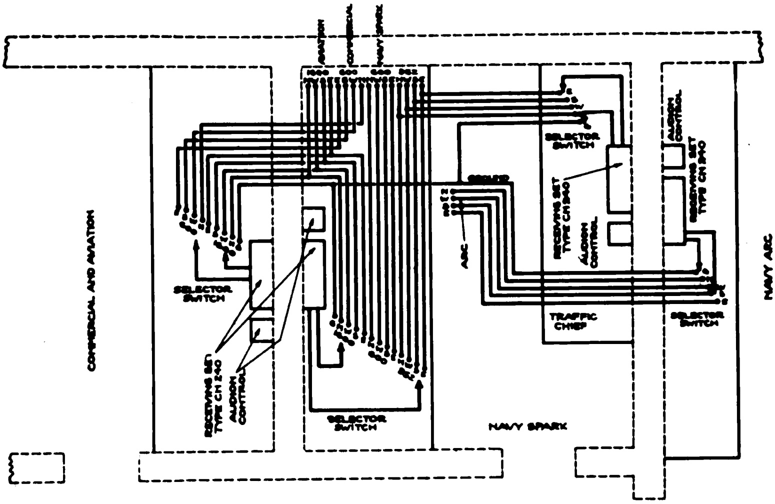

FIGURE 7

There are three receiving rooms, the aviation and commercial, the Navy spark, and the Navy arc. The Navy spark room is situated over the shaft, the aviation and commercial room is west of the Navy spark room, and the Navy arc room is east of the Navy Spark room. Four 600-meter, four 1,600-meter, and one ground wire lead to the selector switch in the aviation and commercial room. Four 600-meter, four 1,600-meter, four 952-meter, and one ground wire lead to the selector switch on the operator's table in the Naval spark room, and from the four 952-meter wires, taps were run to the selector switch on the traffic chief's table in the same room. Four 4,000-meter and one ground wire are led to the selector switch on the table in the Navy arc room. The ground wire referred to in this paragraph consists of a number 4 stranded, rubber-covered cable* which is grounded to the water system and also to a fan net-work of wires under the building. The connections of the receiving apparatus are identical to those used at Great Lakes. Figure 8 shows schematic diagram of receiving system used with the underground wires.

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* Diameter of number 4 wire = 0.204 inch = 0.515 cm.

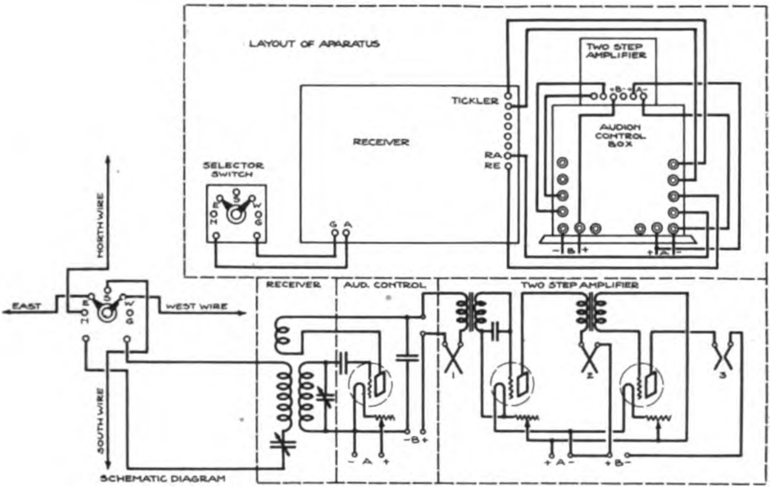

FIGURE 8

The telegraph equipment between the remote control station and the transmitting station, consists of six pairs in lead-covered cable, one pair for each of the two spark transmitting sets, one pair for the arc transmitting set, and one pair for the operator's signal, leaving two spare pairs for future expansion and emergency purposes. At each station all pairs terminate at a ten-pair Western Union button switch, and from this switch leads are run to the various keys and relays. The source of current supply for the telegraph system is obtained from a bank of Edison storage batteries, giving a voltage of 100 volts and a capacity of 37.5 ampere hours. This battery is looped in series with each pair and all pairs are fused with a standard one-ampere telephone fuse. Two motor generator charging sets of 60-volt, 30-ampere capacity are used at the control station for charging the telegraph and audion batteries, one set being used as a spare for emergency purposes. The room in which the motor generators and telegraph batteries and terminal boxes are situated, is electrostatically screened, and is of similar construction to the operating room at the remote control station at Great Lakes.

This screening protects the receiving rooms from stray electric fields from the motor generator charging units. All leads from the charging room to the receiving rooms are run in lead-covered number 14 rubber-covered duplex cable* further to minimize external interference. A schematic diagram showing the telegraph control lines is given in Figure 9.

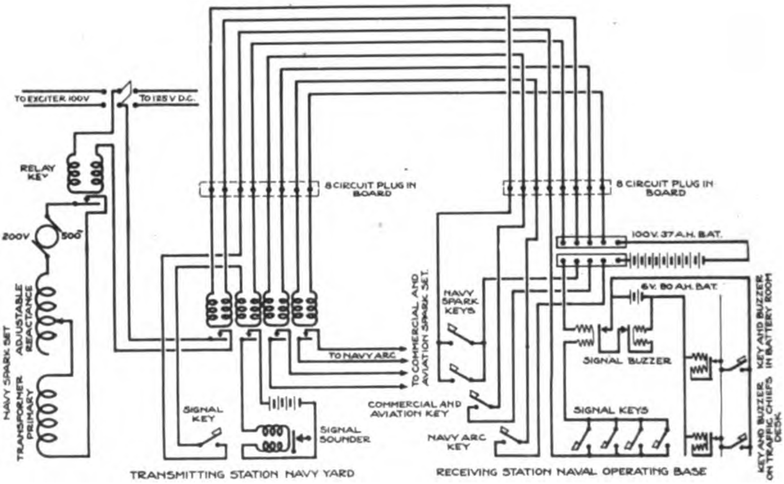

FIGURE 9

REMOTE CONTROL INSTALLATION AT NAVAL RADIO STATION, NEW ORLEANS, LOUISIANA

The Naval radio station at New Orleans is now being equipped with the underground receiving system, by means of which one 5-kilowatt spark and one 30-kilowatt arc are being remotely controlled from a receiving station which is approximately 2,200 feet (671 m.) distant from the center of capacity of the main transmitting antenna. The general scheme as used at New Orleans is similar to that used at the Naval Base at Hampton Roads, with the exception that a double Faraday cage is used to protect the receiving instruments at the receiving station. This double Faraday cage consists of two cages, the outer cage being grounded and the inner cage insulated by means of ebonite or porcelain insulators from the outer cage. These cages are constructed by the use of copper mosquito netting. The double

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* Diameter of number 14 wire = 0.004 inch = 0.162 cm.

Faraday cage gives much better screening effect than a single cage and has been used with great success in foreign countries.

CHARACTERISTICS OF THE UNDERGROUND SYSTEM

The underground system for long wave reception, using optimum length wires, gives roughly the same signal strength as an average 100-foot (31-m.) antenna; while for short wave lengths the signal strength received is a function of the wave length, and it may be said that the shorter the wave length, the weaker the signal. This may be attributed to the skin effect of radio frequency current with reference to the penetrating qualities of such current thru a partial conductor. For a wave length of 600 meters, the signal strength as received by the underground system, is approximately one-twelfth as strong as a signal received with the overhead antenna. For efficient reception on short wave lengths, it requires an amplifier using three stages of radio-frequency, which gives a signal that is approximately of the same strength as that of a signal received on the overhead antenna unamplified.

The question is brought up concerning the efficiency of the system on short wave lengths when using high amplification. To this we may say that reception on short wave lengths by means of the underground system has never been affected by local static. This statement is based upon tests covering a period of two years, wherein the underground system for short wave lengths has been used thru all atmospheric conditions, and at no time have the operators been required to leave the instruments on account of the dangerous effect of local lightning storms. The foregoing statements have been verified at Norfolk, Great Lakes, and New Orleans, where many cases were recorded when the regular antenna had to be grounded to prevent damage to the set, and that during this time the underground system was used exclusively to handle all traffic. Stray elimination by the underground system on long wave lengths is not so pronounced as it is on short wave lengths, but by the use of the underground system in conjunction with the balanced system, this trouble is overcome.

The underground system has excellent directional qualities. Using two wires in the direction of the transmitting station, the maximum signal is obtained; while using two wires which lie in a direction which is at right angles to the transmitting station, the minimum signal is obtained, and in many cases the signal is not heard. It is possible that the operator on watch can, at any time, ascertain the approximate direction from which the signals are coming by means of using various combinations of wires. A good stand-by tune, by which the operator can hear all stations, is obtained by the use of two wires at right angles to each other, namely an east-and-south or east-and-north combination, and so on.

Because of the selectivity of the system and the sharp tuning of the primary, the system is admirably adapted to distant control work. Using a comparatively tight coupling at the remote control station, Naval Base, Norfolk, no difficulty is had in reading stations with one hundred meters difference in wave length when Norfolk is transmitting. The Norfolk Navy Yard being due south of the distant control station, ships operating on 952 meters at sea, east of the remote control station, are readable thru Norfolk when both stations are transmitting on 952 meters. Many combinations are obtained by using different sets of wires, which help out the operator in eliminating interfering stations.

It is very essential in this system that all wires be perfectly insulated; wires that are grounded bring in more strays altho when wires have been perfectly grounded, the stray ratio is still equal to and often better than that of the regular antenna. However, to obtain the best results from the system, the wires should be clear of all grounds. Grounded wires do not appreciably affect the received signal strength. This, of course, depends upon the conductivity of the medium in which these wires are buried, the greater the conductivity, the weaker the signal.

For short wave lengths it is highly essential that the optimum length of wire for each wave length be used. Altho by means of high amplification the proper signal strength may be obtained, for efficient results the optimum length should be used. For long wave length reception it has been found that a length of 1,500 feet (458 m.) of wire is capable of receiving signals up to 17,000 meters wave length, but where available space can be obtained, it will be of great advantage to use wires as near optimum length as possible, as by the use of these wires, minimum amplification is needed, and therefore troubles experienced from internal amplifier noises are reduced to a minimum. Referring to a previous statement in this paper, with reference to the depth to which wires should be buried to obtain proper water level, it may be stated that there are many cases where the cost of digging trenches to the proper water level will be prohibitive. In this case it may be said that excellent results have been obtained at various stations, where wires have been buried from two to four feet (0.6 to 1.2 m.) below the surface. As regards stations situated along the seacoast, it may be stated that it is highly detrimental to the efficiency of the station if short wave length wires are buried or submerged in salt water or salt water marshes, whereby these wires will be at times covered to a certain depth with salt water. The salt water being many times more conductive than fresh water, absorbs to a great extent short wave length signals, while for long wave lengths this effect is not so pronounced. For long wave lengths the depth of penetration is a function of the wave length, and the longer the wave length the greater the penetration. In the case of installing a system along the seacoast, it is recommended that the short wave lengths wires be buried in moist ground and not in salt water marshes. The case is opposite where fresh water lakes or rivers are near the site of the receiving station, for in this case it has been found that wires submerged in fresh water lakes or rivers give much better results than wires buried in moist ground or salt water marshes.

There is a tendency on the part of operators to disparage the ground wire system during the winter months when strays are at a minimum, because of the fact that the signals are so much weaker than they are on an antenna with the same degree of amplification, but during the summer months the ground wire system demonstrates its superiority. In the matter of eliminating interference from nearby stations it is, of course, always superior. The first short wave receiving sets supplied for ground wire work were decidedly inefficient, and for this reason short wave signals were not what they should have been. This difficulty is being remedied at the present time by issuing, for short wave work, highly efficient regenerative receivers with both radio and audio frequency amplification. Preliminary tests of these receivers have shown that they are peculiarly well adapted for this work and no further difficulties in short wave reception on ground wires are anticipated.

SUMMARY

1. The general theory of remote control using the ground wire installation for receiving has been discussed.
2. The installation at Great Lakes has been described in detail.
3. The installation at Norfolk has been described in detail.

4. The work under way at New Orleans has been briefly mentioned.

5. The general peculiarities of the system in actual operation have been commented upon.

SUMMARY: By utilizing the directional selectivity of subterranean receiving antennas, it has proven possible to install receiving stations in the immediate vicinity of transmitting stations.

The proper design of the underground antennas, the construction of the shielded rooms for the receiving sets, and the mode of "remotely" controlling the nearby transmitting station are described in detail.

The complete design, lay-out, and installation of several such Naval "remote" control stations is then given. Their proper mode of use and advantages are considered.

DISCUSSION

John Mills: Commander Taylor in the present paper and in several recent and interesting papers, has described methods of reducing static strays and also foreign station interference. In some instances he has described his results in terms of a ratio giving the advantage of one system over another. The subject is one of very great interest. In his own papers and in others which have recently appeared, the results of work were given in a form which I am unable to interpret quantitatively. I wish, therefore, to suggest for the consideration of the Institute a method for the quantitative comparison of receiving systems so far as concerns their ability to reduce static or interference. As it happens, the method which I am suggesting is one which I devised in the early part of 1916 when, for two or three months, I was employed in studying receiving circuits.

It is first necessary to choose arbitrarily some simple receiving circuit and antenna as a standard of comparison to which all other forms of receiving circuit or antennas may be referred. A switching system is provided so that one may listen alternately on the standard system and on the one under examination. Tests should be made by reading signals in code, not ordinary English words, for two minute intervals, first on one system and then on the other over a considerable period of time. The tests should be made not only by good operators and those specially gifted with the sense of absolute pitch, but by mediocre and even poor operators. The signals should be locally generated and supplied to the antenna thru some coupling arrangement, either a loose inductive coupling or a unilateral device like a vacuum tube. These signals are superimposed upon the static or interference. The intensity of the signals should be adjusted to be equal in the two systems as observed in the telephones. Their intensities should also be so adjusted with reference to the static or interference that the latter are appreciably troublesome. As a rough measure of the advantage of the system under examination, we then have the ratio of the percent of correct signals received thru it to the corresponding number received on the standard system. Let these percentages be N_2 and N_1, respectively.

To obtain a more exact and physical measure of the advantages of one system over the other, I would proceed as follows: Work entirely with the standard system and so adjust the current, energy, voltage or audibility of the signal, whichever one wishes, that the percentage of intelligible signals is N_1. Now alter the intensity of the locally generated signals until the percent of correctly received signals is altered to a second value of N_2. Personally I prefer to deal with the energies which are supplied in the two cases, that is, the square of the input currents rather than with current, voltage or audibility. Let the energies corresponding to the conditions of the receiving set, which have just been described, be respectively, E_1 and E_2. Then the ratio of $E_2 : E_1$ gives a quantitative expression for the efficiency of system number 2 as compared to the standard system.

It is evident that as long as the type of interference and the intensity of interference does not change, there will be a fairly definite relation between the percentage of intelligibility and the energy of the local signal in the standard circuit. Under certain conditions, therefore, it may not be necessary to form the calibration described above, for all of the conditions made within the circuits under test, since the relation of N and E, just mentioned, can be determined by a few observations.

It is also evident that the system which is described permits of a very satisfactory quantitative measure of static or station interference. It does not, of course, give root-mean-square or peak value of static, which are probably only of academic interest, but it does give a measure of what the engineer is usually interested in, namely, the difficulty of getting a message thru. Using therefore, a calibrated detector and a calibrated oscillator as a source of local signals, it is possible to compare static or station interference on different occasions by the inverse ratio of the percentages of correct signals which are received on the two different occasions, assuming, of course, the same efficiency of detection and also of strength of local signal.

I make no claim, of course, that this is the optimum system for the quantitative comparison of receiving systems or static and station interference. I feel, however, that so much is now being written and published on these matters that it would be of great advantage to the Institute to consider seriously the adoption of some simple and standard method for expressing the results obtained.

THE ROME RADIO STATION OF THE ITALIAN NAVY*

By

Cyril F. Elwell

(Radio Engineer, Paris, France)

The design of a high-power radio plant under the stress of war conditions tends to emphasize factors which would not have to be considered in peace time. It may be of interest to review the reasoning which resulted in the design hereafter described. Naturally, the outstanding factor was that of time, as the plant was required because of the congestion of both the cable and radio equipment to America, as well as to the Italian colonies in Africa.

ANTENNA SUPPORTS

SELECTION OF TYPE

Of the three available forms of antenna support, namely: self-supporting steel towers, guyed steel masts, and guyed wooden masts, the first was immediately rejected as not being compatible with the essential factor of time. The choice then lay between steel and wood. The distance over which good communication was desired being 4,200 miles (7,000 km.), a large effective height was a necessity. For the same height of mast, wood gives an effective height appreciably superior to steel. Altho the highest existing wooden masts were 608 ft. (185 m.) high, it was considered feasible to adopt the same type of construction for masts of 714 ft. (218 m.) in height, without having recourse to excessively large timbers in the bottom sections. To obtain the same effective height with steel, it would have been necessary to erect masts of at least 750 ft. (228 m.), which with the lightest design would have entailed the employment of about 300 tons (27,000 kg.) of fabricated steel for three masts. With demands much above possible production, the factors of price and time of delivery were such as made the decision in favor of 105,000 board feet (248 cu. m.) of pitch pine an easy matter.

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* Received by the Editor, October 14, 1919.

DESIGN DETAILS

Wind load employed for the calculation of the stresses in the guys, bracing and columns—61.5 lb./ft.^{2} (300 kg./m.^{2}).

Horizontal component of the antenna pull—22,000 lb. (10,000 kg.).

Factor of safety allowed for the guys—varied from 3 to 4, as a single size of wire rope was used for simplicity.

Maximum stress allowed in the wood—1,000 lb./in.^{2} (70 kg./cm.^{2}).

Number of sections—eleven.

Longest unsupported section—72 ft. (22 m.).

Number of supporting guys—11 sets of three at 120°.

Longest section of guy—between porcelain insulating breaks —82 ft. (25 m.).

Size of bottom column—12.6 in.×16 in. (32×40 cm.).

Size of top columns—12.6 in.×7.1 in. (32×18 cm.).

Figure 1 shows the bottom of a mast in detail. Figure 2 shows one complete mast, and Figure 3 the complete station.

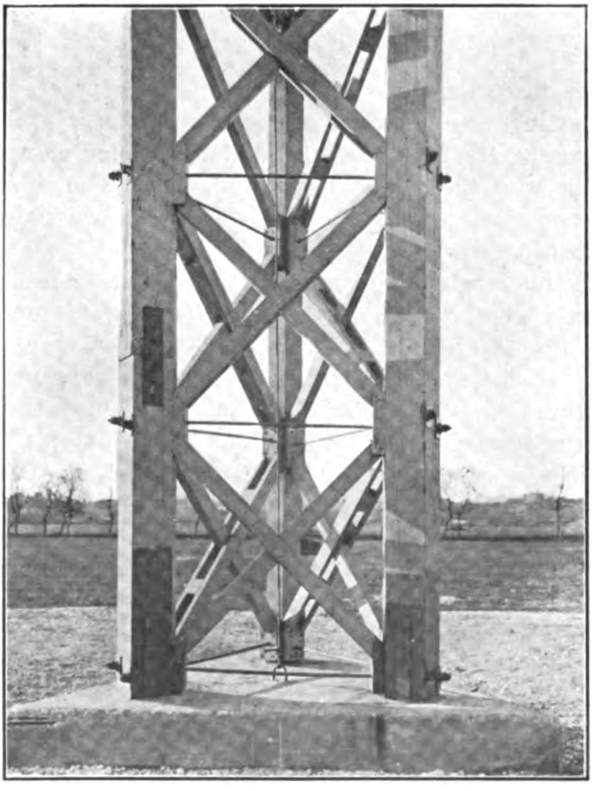

FIGURE 1—Foot of Wooden Tower

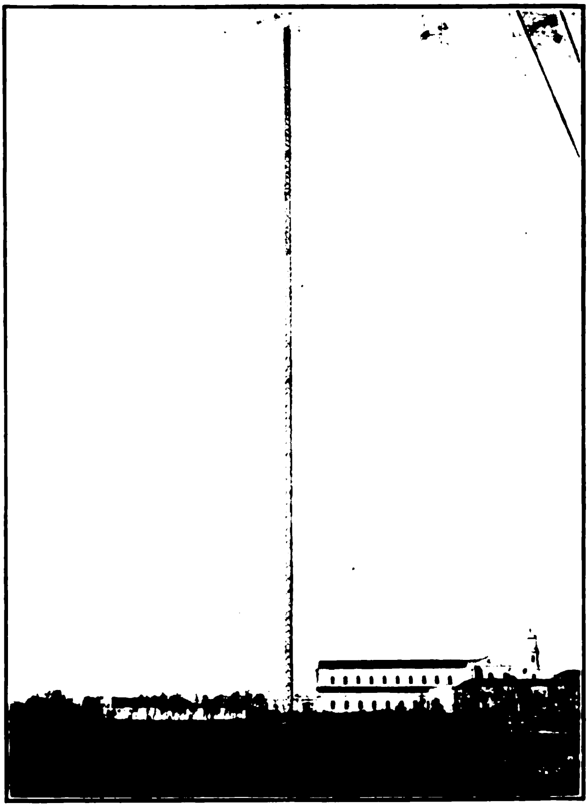

FIGURE 2—Wooden Mast 714 Feet (218 Meters) High

ANTENNA

As large an antenna capacity as possible being required, it was considered feasible to place three masts at the corners of an equilateral triangle with sides of 984 ft. (300 m.), without exceeding the horizontal pull of 22,000 lb. (10,000 kg.) allowed in the design of the masts.

The triangular form of antenna, with only three strings of insulators to support it, is as close to the ideal as can be realized. The arc generator is very sensitive to poor antenna insulation, and an antenna with a minimum number of strings of insulators to support it, is always to be preferred.

The individual antenna wires were cut so that the sag in the messenger cables (guys between towers) was accentuated in the horizontal plane, in order to avoid too large a sag, and consequent loss of effective height, in the vertical plane.

The capacity of the antenna is 0.011 mfd.

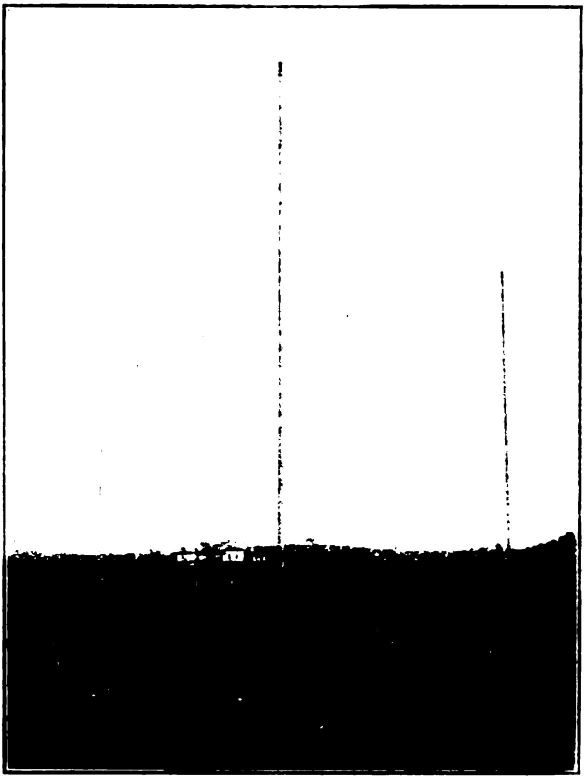

Figure 3—Radio Station at San Paolo, Rome, Italy

Ground Connection

The usual radiating network of buried copper wires covering an area greater than the antenna was employed. This was supplemented by means of a number of wells down to ground water.

Radio Apparatus

The arc generator rated at 200 kilowatts (250 amperes at 800 volts), installed in duplicate, is shown in Figure 4. Its weight, without concrete support, is 3.75 tons (3,750 kg.).

The control switchboard consists of five marble panels. Two panels for each arc generator and a change-over panel for the dynamo protection coils and the respective anodes.

The series resistance for starting the arc is reduced to zero by means of five electromagnetic contactors controlled by means of a master switch mounted on the control panel.

The water supply for cooling the anode, cathode, chamber, lid, and other parts of the arc generator, is electrically interlocked with the 750-volt direct current supply.

The sending inductance is wound with 1.5 in. (38 mm.) diameter copper tubing. Its maximum valve is 2,500 microhenrys.

The choke coils for the protection of the d. c. dynamo, each with an inductance of 5,000 microhenrys, being wound with flat copper strip.

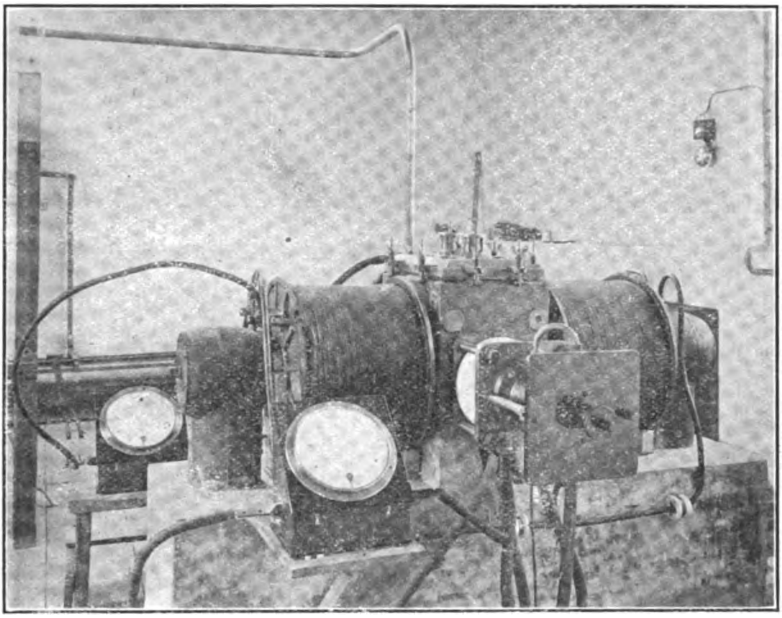

FIGURE 4—200-Kilowatt Arc at Rome Radio Station

The motor-generator set, shown in Figure 5, consists of a 3-phase, 42-period, 500-volt motor, direct coupled to a 350-ampere, 750-volt, compound wound, interpole dynamo made by Ercole, Marelli and Company, of Milan.

For over a year this station has played a by no means unimportant part in the fight for Liberty which has just been brought to the only issue possible for the peace and progress of the world.

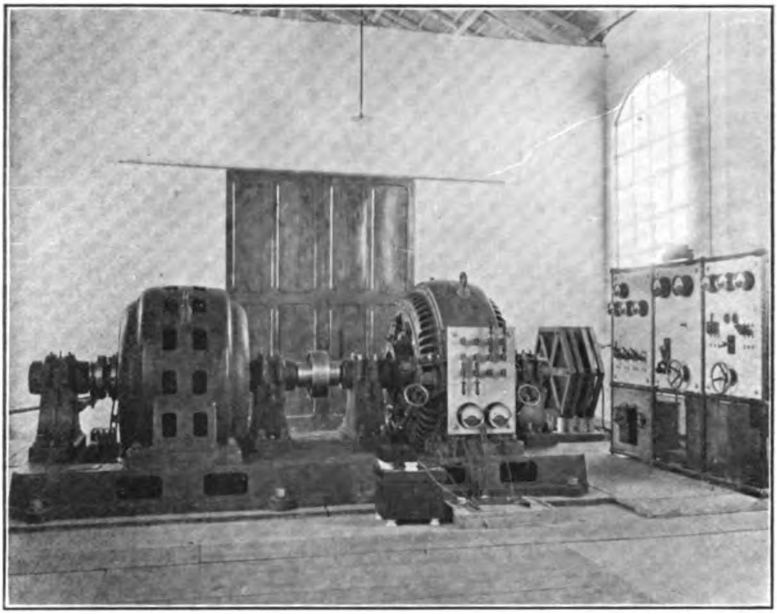

Figure 5—Motor Generator and Control Panels, Rome Radio Station

I take this opportunity to thank Admirals Pinelli and Simion, Commandants Micchiardi and Pession, Professor Vallauri and Mr. S. P. Wing for their invaluable help and co-operation.

SUMMARY: **The antenna, ground system, arc, generating and control equipment at the Rome transmitting station are described with numerical data.**

SIMULTANEOUS TRANSMISSION AND RECEPTION IN RADIO TELEPHONY*

By

Noboru Marumo

(Imperial Japanese Electro-Technical Laboratory, Tokyo, Japan)

Introduction

In spite of the various devices which have been tried to minimize induction into the receiving system from its own transmitter, the effect has generally been too strong to permit maintaining the detector in sensitive adjustment, and it has been practically impossible to receive signals while the nearby transmitter was radiating. Even tho many inventors have tried to overcome this difficulty, and to succeed in sending and receiving simultaneously in radio telegraphy and telephony, the only practical duplex radio telegraph system seems to have been that of the Marconi Company using their well-known antenna arrangement, as shown in Figure 1. This requires sending and receiving stations several miles apart, and is therefore not suitable for adoption in small radio telephonic stations, and particularly not in ship stations. However, the recent development of the vacuum

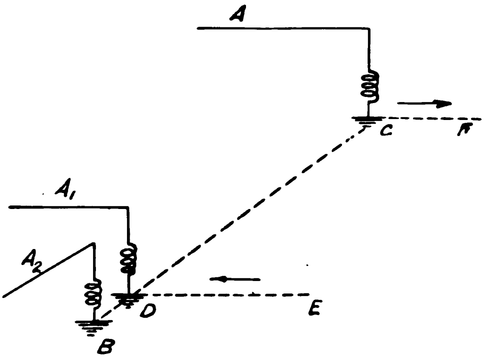

Figure 1

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* Received by the Editor, April 7, 1919.

tube has improved conditions considerably, and we have succeeded in permanently connecting receiving and transmitting circuits, even to the same antenna, without objectionable effects.

Dr. Wichi Torikata and his staff in the Imperial Japanese Electro-Technical Laboratory, Tokyo, including the writer, have published an account of the practical and successful results obtained by duplex or simultaneous sending and receiving in radio telephony, in the "Proceedings of the Electrical Engineers' Society, Japan," August 31, 1917. The connections used are shown in the diagram of Figure 2. It is based on the use of the vacuum tube as detector as well as oscillator, on the sending and receiving circuits being permanently connected to the same antenna, and on minimizing induction to the receiving bulb by properly coupling the balancing coil L_2'' or the absorbing circuit *W* to the receiving set primary or secondary.

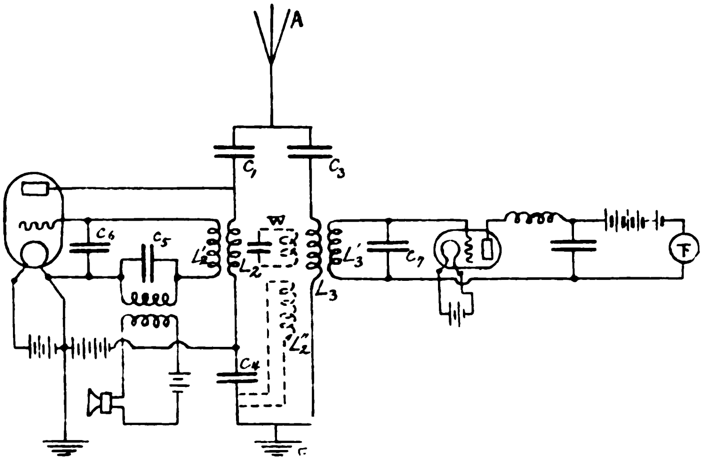

FIGURE 2

So far as the writer is aware, a list of all descriptions previously published dealing with duplex or simultaneous sending and receiving systems in radio telephony follows*.

(1) Device patented by Marconi Company[1]. Figure 1, published 1911.

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*[1] British patent to G. Marconi, and Marconi's Wireless Telegraph Company, number 13,020 of May 30, 1911.

(2) Device patented by Yokoyama[2]. Figure 3, published 1913.

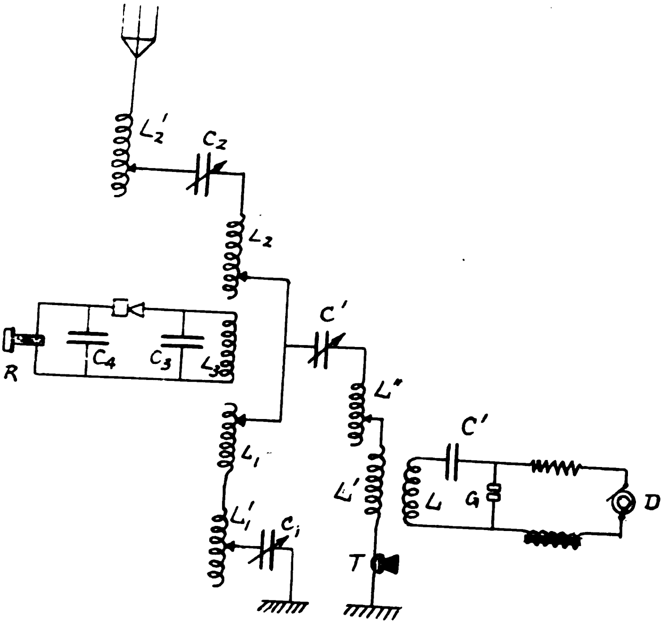

FIGURE 3

(3) Device published by Torikata .[3] Figure 4, published 1913.

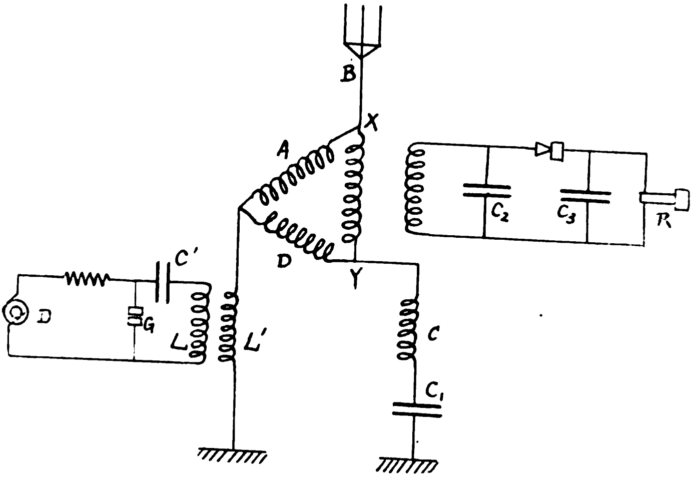

FIGURE 4

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[2] Japanese patent to E. Yokoyama, number 25,140 of December 17 ,1913.

[3] M. Kitamura, "Dainibu Kenyu Kaishi," published by Imperial Japanese Electro-Technical Laboratory, Tokyo, in April, 1913.

(4) Device patented by Fessenden.[4] Figure 5, published 1916.

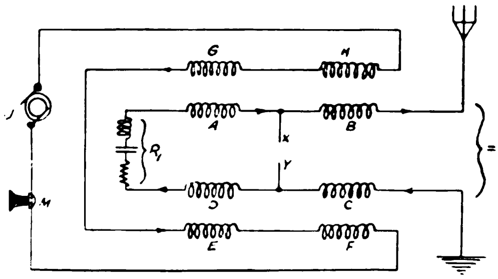

FIGURE 5

(5) Device patented by Carson.[5] Figure 6, published 1916.

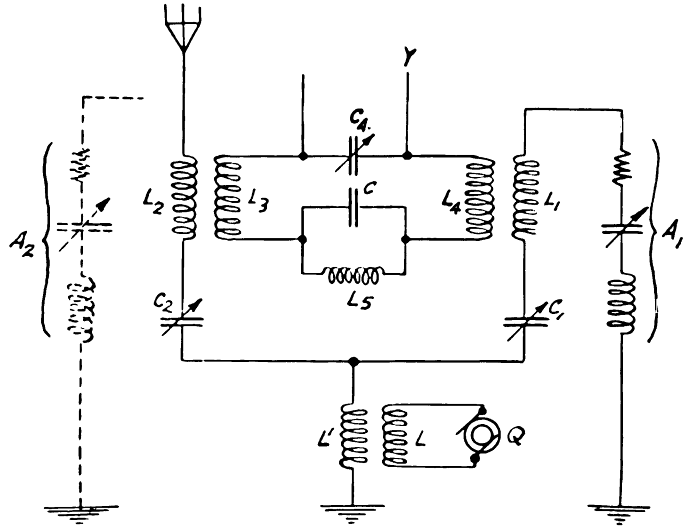

FIGURE 6

(6) Device published by Torikata and his staff.[6] Figure 2, published 1917.

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[4] U. S. Patent to R. A. Fessenden, number 1,170,969 of February 8, 1916.

[5] U. S. Patent to J. R. Carson, number 1,188,531 of June 27, 1916.

[6] W. Torikata, "Proceedings of Electrical Engineers' Society, Japan," number 349, August 31, 1917.

(7) Device published by Saeki.[7] Figure 7, published 1917.

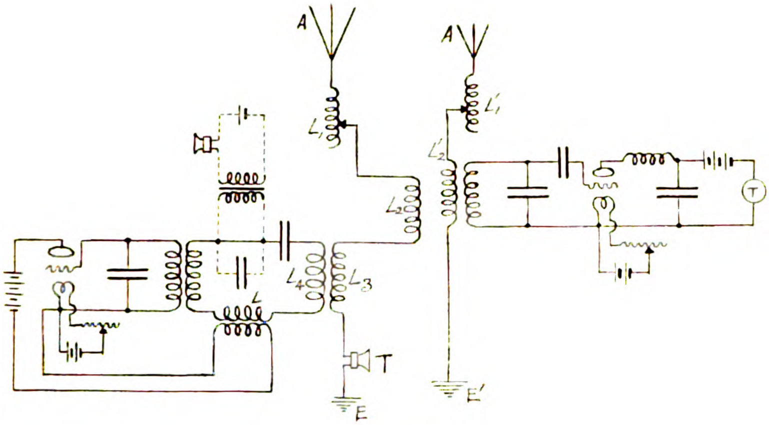

FIGURE 7

(8) Device patented by Englund.[8] Figure 8, published 1917, 1918.

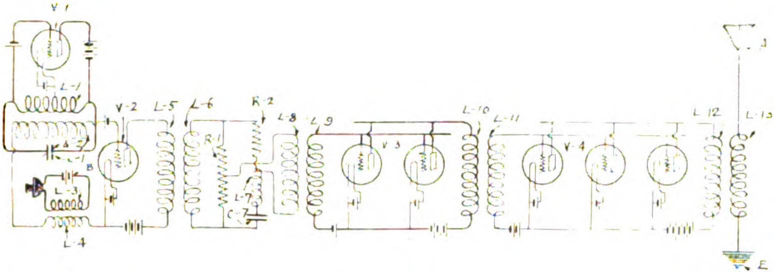

FIGURE 8

(9) Device patented by Espenschied.[9] Figures 9a and Figure 9b, published 1918.

As regards the practical utility of these devices, the writer has no detailed information as to the European and American devices, but the device described by Dr. Torikata and his staff, as shown in Figure 2, is quite successful in actual communication, enabling, for example, persons on ships in the harbor to speak directly with any city telephone subscriber. Moreover, the idea has been extended to permit considerable utilization of "wired wireless" (guided radio communication) over ordinary telephone and telegraph lines and also over power transmission lines. It is therefore deemed desirable to give the principles of the idea and data on its utility in the following pages.

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[7] M. Saeki, "Proceedings of Telegraph and Telephone Engineers' Society, Japan," number 4, November 20, 1917.

[8] U. S. Patent to C. R. Englund, number 1,245,446 of November, 1917, and number 1,258,548 of March, 1918.

[9] U. S. Patent to L. Espenschied, number 1,256,889 of February 19, 1918.

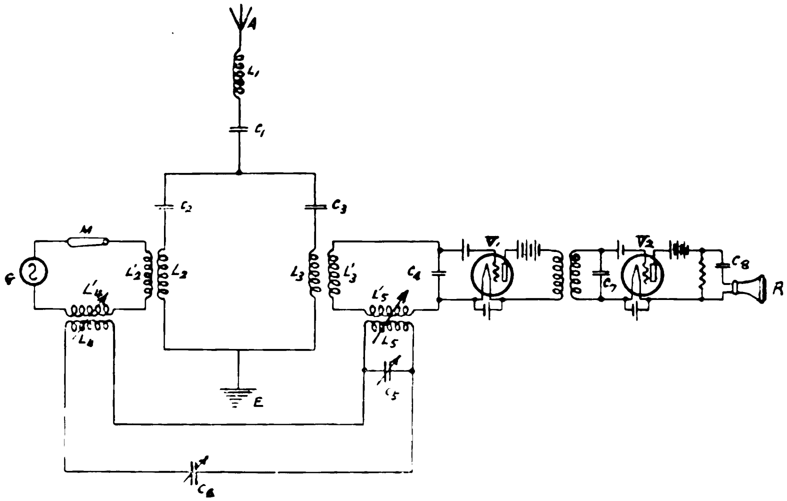

FIGURE 9A

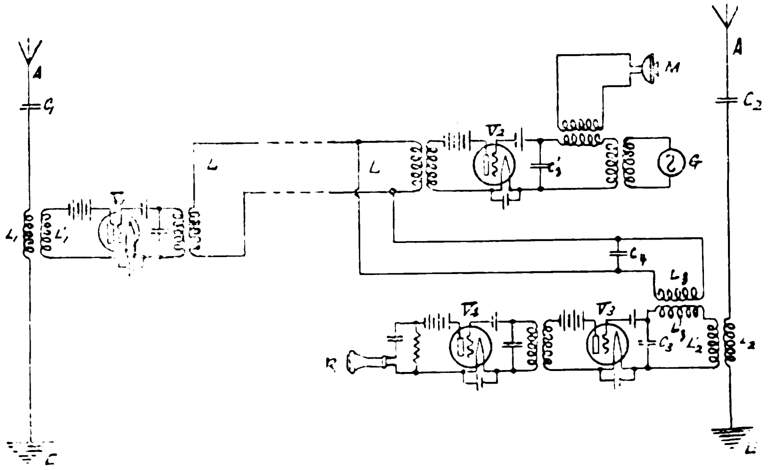

FIGURE 9B

2. Some Requirements for Simultaneous Transmission and Reception

On putting the simultaneous transmission and reception in radio communication into practice, great difficulty was experienced in excluding induction into the receiving circuit from its own transmitter, as mentioned above. The methods devised by many investigators, such as those described in the preceding articles, can be considered to be based on one or more of the following principles:

a. Using separate sending and receiving antennas
b. Using different wave lengths
c. Using differential or bridging circuits
d. Using balancing-out or neutralization circuits
e. Using voice-controlled switching arrangements.

Systems based on these principles may be feasible or valuable or neither, depending on conditions at the station. The experience of the writer indicates that the arrangement of Figure 10 is very convenient in practice for radio telephony, as is also the arrangement of Figure 11 for the "wired wireless" (guided radio) telephony.

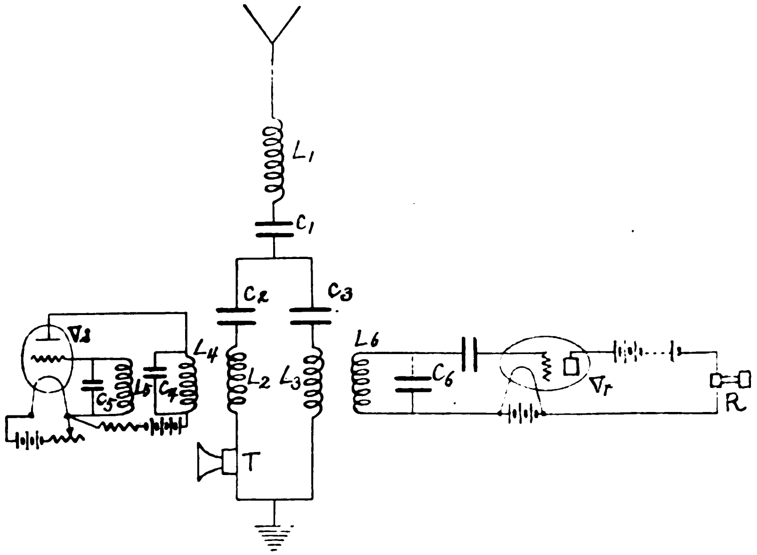

Figure 10

The behavior and basic principles of the circuits shown in Figure 10 and Figure 11 are explained below.

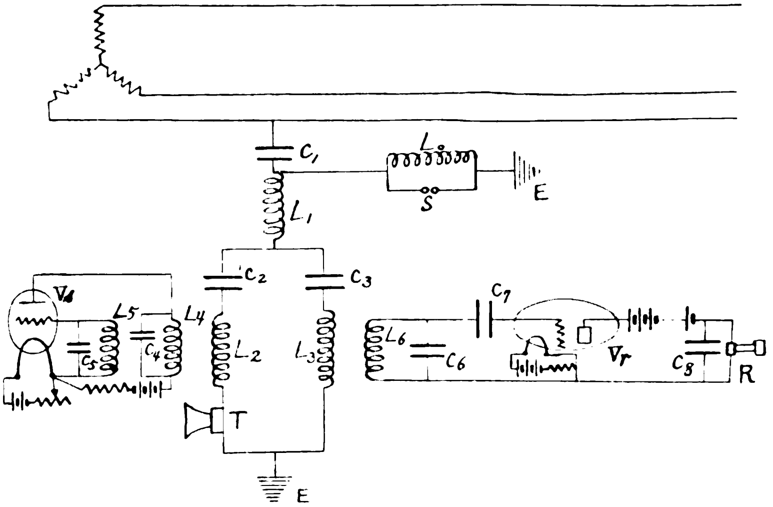

FIGURE 11

3. GENERAL THEORY

In Figure 12, the current in the antenna, as well as that in the sending and receiving coils, may not be constant thruout the circuit, but may have a variable distribution in the circuit. The antenna capacity is distributed along its wires. However, to simplify the problem, we assume the current at any point A in the circuit to be constant thruout the circuit, as indicated in Figures 12 and 13, and we shall call the equivalent capacity, inductance, and resistance of the antenna, sending, and receiving circuits respectively C_1, L_1, R_1; C_2, L_2, R_2; and C_3, L_3, R_3, as indicated in Figure 13.

Suppose now an undamped oscillation $e_1 = E_1\varepsilon^{j\omega_1 t}$ is induced in the circuit L_1C_1 by the incoming signals. Then we have

$$\left(L_1\frac{di_1}{dt}+R_1i_1+\frac{1}{C_1}\int i_1\,dt\right)-\left(L_2\frac{di_2}{dt}+R_2i_2+\frac{1}{C_2}\int i_2\,dt\right)=E\varepsilon^{j\omega_2 t} \quad (1)$$

$$L_2\frac{di_2}{dt}+R_2i_2+\frac{1}{C_2}\int i_2\,dt=L_3\frac{di_3}{dt}+R_3i_3+\frac{1}{C_3}\int i_3\,dt \quad (2)$$

$$i_1+i_2+i_3=0 \quad (3)$$

Putting

$$i_1 = A_1 \varepsilon^{j\omega_1 t}, \quad i_2 = A_2 \varepsilon^{j\omega_1 t}, \quad \text{and} \quad i_3 = A_3 \varepsilon^{j\omega_1 t} \tag{4}$$

where A_1, A_2, and A_3 are certain complex quantities, we have

$$\left.\begin{aligned} &A\left\{R_1 + j\left(\omega_1 L_1 - \frac{1}{\omega_1 C_1}\right)\right\} - A_2\left\{R_2 + j\left(\omega_1 L_2 - \frac{1}{\omega_1 C_2}\right)\right\} = E \\ &A_2\left\{R_2 + j\left(\omega_1 L_2 - \frac{1}{\omega_1 C_2}\right)\right\} = A_3\left\{R_3 + j\left(\omega_1 L_3 - \frac{1}{\omega_1 C_3}\right)\right\} \\ &A_1 + A_2 + A_3 = 0. \end{aligned}\right\} \tag{5}$$

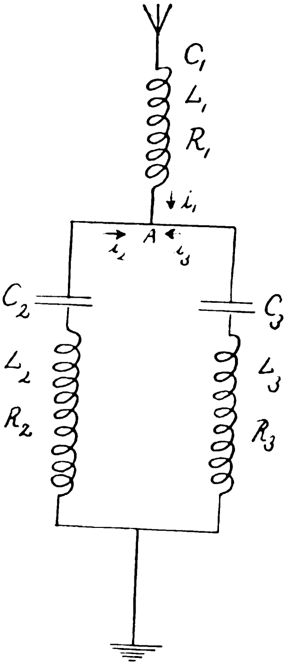

FIGURE 12

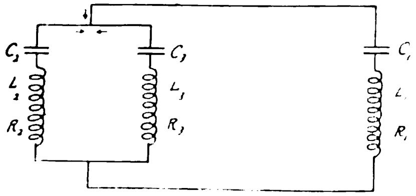

FIGURE 13

[illegible]

[illegible]

[illegible]

[illegible]

we have, from equation (5),

$$\left.\begin{aligned}
A_1 &= \frac{\{(R_2+R_3)+j(S_2+S_3)\}E}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1S_2+S_2S_3+S_3S_1)\}+j\{R_1(S_2+S_3)+R_2(S_3+S_1)+R_3(S_1+S_2)\}}\\
A_2 &= \frac{-(R_3+jS_3)E}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1S_2+S_2S_3+S_3S_1)\}+j\{R_1(S_2+S_3)+R_2(S_3+S_1)+R_3(S_1+S_2)\}}\\
A_3 &= \frac{-(R_2+jS_2)E}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1S_2+S_2S_3+S_3S_1)\}+j\{R_1(S_2+S_3)+R_2(S_3+S_1)+R_3(S_1+S_2)\}}
\end{aligned}\right\}\quad(7)$$

Similarly when the electromotive force $e_2=E_2\varepsilon^{j\omega_2t}$ is induced in the circuit, we have

$$\left.\begin{aligned}
A_1' &= \frac{-(R_3+jS_3')E_2}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1'S_2'+S_2'S_3'+S_3'S_1')\}+j\{R_1(S_2'+S_3')+R_2(S_3'+S_1')+R_3(S_1'+S_2')\}}\\
A_2' &= \frac{\{(R_1+R_3)+j(S_1'+S_3'\}E_2}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1'S_2'+S_2'S_3'+S_3'S_1')\}+j\{R_1(S_2'+S_3')+R_2(S_3'+S_1')+R_3(S_1'+S_2')\}}\\
A_3' &= \frac{-(R_1+jS_1')E_2}{\{(R_1R_2+R_2R_3+R_3R_1)-(S_1'S_2'+S_2'S_3'+S_3'S_1')\}+j\{R_1(S_2'+S_3')+R_2(S_3'+S_1')+R_3(S_1'+S_2')\}}
\end{aligned}\right\}\quad(8)$$

Expressing the amplitude of A_1, A_2, A_3, and A_1', A_2', A_3' respectively with $[A_1]$, $[A_2]$, $[A_3]$, and $[A_1']$, $[A_2']$, $[A_3']$, we have

$$[A_1]:[A_2]:[A_3]::\sqrt{(R_2+R_3)^2+(S_2+S_3)^2}:\sqrt{R_3^2+S_3^2}:\sqrt{R_2^2+S_2^2} \qquad (9)$$

$$[A_1']:[A_2']:[A_3']::\sqrt{R_3^2+S_3'^2}:\sqrt{(R_1+R_3)^2+(S_1'+S_3')^2)}:\sqrt{R_1^2+S_1'^2} \qquad (10)$$

$$\frac{[A_3]}{[A_2]}=\frac{\sqrt{R_2^2+S_2^2}}{\sqrt{R_3^2+S_3^2}} \quad \text{in receiving circuit} \qquad (11)$$

$$\frac{[A_1']}{[A_3']}=\frac{\sqrt{R_3^2+S_3'^2}}{\sqrt{R_1^2+S_1'^2}} \quad \text{in sending circuit} \qquad (12)$$

That is, the receiving current divides between the receiver and the transmitter circuits in inverse proportion to their impedances, as also does the transmitting current between the antenna and receiving circuits.

As for the natural frequency of the system, putting $E_1=0$ in equation (5) and eliminating A_1, A_2, and A_3, we have

$$\{(R_1R_2+R_2R_3+R_1R_3)-(S_1S_2+S_2S_3+S_3S_1)\}+j\{R_1(S_2+S_3)+R_2(S_1+S_3)+R_3(S_2+S_1)\}=0 \qquad (13)$$

The equation is of the second order in ω_1, and therefore we have

$$\left.\begin{aligned}\omega_1&=\omega_3+j\,a_1\\ \omega_1&=\omega_4+j\,a_2\end{aligned}\right\} \qquad (14)$$

where ω_3, ω_4 and a_1, a_2 are functions of R_1, R_2, R_3, L_1, L_2, L_3, and C_1, C_2, C_3. Assuming R_1, R_2, and R_3 to be small and neglecting them, we have, from (13),

$$\left.\begin{aligned}&S_1 S_2+S_2 S_3+S_3 S_1=0\\&\left(\omega_1 L_1-\frac{1}{\omega_1 C_1}\right)\left(\omega_1 L_2-\frac{1}{\omega_1 C_2}\right)+\left(\omega_1 L_2-\frac{1}{\omega_1 C_2}\right)\left(\omega_1 L_3-\frac{1}{\omega_1 C_3}\right)+\left(\omega_1 L_1-\frac{1}{\omega_1 C_1}\right)\left(\omega_1 L_3-\frac{1}{\omega_1 C_3}\right)=0\end{aligned}\right\} \quad (15)$$

$$\therefore \omega_{3,4}=2\pi f_{3,4}=\sqrt{\frac{\left\{\frac{1}{C_1}(L_2+L_3)+\frac{1}{C_2}(L_1+L_3)+\frac{1}{C_3}(L_2+L_1)\right\}\pm\sqrt{\left\{\frac{1}{C_1}(L_2+L_3)+\frac{1}{C_2}(L_1+L_3)+\frac{1}{C_3}(L_1+L_2)\right\}^2-4\,(L_1 L_2+L_2 L_3+L_3 L_1)\left(\frac{1}{C_1 C_2}+\frac{1}{C_2 C_3}+\frac{1}{C_3 C_1}\right)}}{2\,(L_1 L_2+L_1 L_3+L_3 L_1)}} \quad (16)$$

This is the formula obtained by W. H. Eccles and A. J. Makower[10].

[10] W. H. Eccles and A. J. Makower: "The Electrician," volume 75, number 25, page 905, September 24, 1915.

4. BEST CONDITIONS FOR THE SIMULTANEOUS TRANSMISSION AND RECEPTION

The adjustment for minimizing the current flowing in the receiving system due the sending current is essential in successful simultaneous transmission and reception. In the arrangements shown in Figures 10 and 11, the best result will be obtained by the following adjustment:

(1) Tune the sending wave length to one of the natural wave lengths of the system, say λ_1, and the receiving wave length to the other, say λ_2.

(2) Adjust the circuits to minimize the current flowing in the receiving branch due to the sending current by making the ratio $\frac{[A_1']}{[A_3']}=\frac{\sqrt{R_3^2+S_3'^2}}{\sqrt{R_1^2+S_1'^2}}$ a maximum.

(3) Adjust the circuits to make the ratio $\frac{[A_3]}{[A_2]}=\frac{\sqrt{R_2^2+S_2^2}}{\sqrt{R_3^2+S_3^2}}$ as large as possible, thus improving the receiving efficiency while maintaining the above conditions, (1) and (2).

It is seen that the second condition will satisfy also the requirement of adjusting the circuits to get the maximum transmitting efficiency.

For the second condition above, remembering equation (12), we have:

$$\frac{[A_1']}{[A_3']}=\frac{\sqrt{R_3^2+S_3'^2}}{\sqrt{R_1^2+S_1'^2}}=\frac{\sqrt{R_3^2+\left(\omega_2 L_3-\frac{1}{\omega_2 C_1}\right)^2}}{\sqrt{R_1^2+\left(\omega_2 L_1-\frac{1}{\omega_2 C_1}\right)^2}}$$

where $\frac{\omega_2}{2\pi}$ is the frequency of the transmitting current oscillation. And when $R_1 \pm R_3 \neq 0$, $\omega_2 L_3-\frac{1}{\omega_2 C_3}\neq 0$, the second condition must fulfil the next equation,

$$\left.\begin{aligned}\omega_2 L_1-\frac{1}{\omega_2 C_1}&=0\\ \omega_2^2=\frac{1}{C_1 L_1}&\end{aligned}\right\}\qquad(17)$$

From the first condition, $\frac{\omega_2}{2\pi}$ must be one of the natural frequencies of the circuit, say $\frac{\omega_3}{2\pi}$, that is, it must satisfy equation (15). Therefore we have:

$$\left.\begin{aligned} C_1 L_1 &= C_2 L_2 \\ \omega_2^2 &= \frac{1}{C_1 L_1} = \frac{1}{C_2 L_2} \end{aligned}\right\} \quad (18)$$

From (15) and (18), we have, as regards the other natural frequency $\frac{\omega_4}{2\pi}$ which must be tuned to the received current frequency:

$$\left(\omega_4 L_1 - \frac{1}{\omega_4 C_1}\right)\left(\omega_4 L_2 - \frac{1}{\omega_4 C_2}\right) + \left(\omega_4 L_2 - \frac{1}{\omega_4 C_2}\right)\left(\omega_4 L_3 - \frac{1}{\omega_4 C_3}\right) + \left(\omega_4 L_1 - \frac{1}{\omega_4 C_1}\right)\left(\omega_4 L_3 - \frac{1}{\omega_4 C_3}\right) = 0 \quad (19)$$

and

$$\left.\begin{aligned} &\frac{1}{C_1 L_1} = \frac{1}{C_2 L_2}, \\ &\omega_4^2 = \frac{\frac{L_2}{C_1} + \frac{1}{C_3}(L_1 + L_2)}{L_1 L_2 + L_2 L_3 + L_3 L_1} \end{aligned}\right\} \quad (20)$$

That is, we have to tune the natural frequency of the antenna circuit relative to that of the sending circuit so as to make the transmitting efficiency a maximum or the branch current to the receiver circuit a minimum, thus disturbing the receiver bulb as little as possible.

Again, from (18) and (20), we have—for transmitter current:

$$\frac{[A_1']}{[A_3']} = \frac{\sqrt{R_3^2 + \left(\omega_2 L_3 - \frac{1}{\omega_2 C_3}\right)^2}}{R_1} \quad (21)$$

For receiver current

$$\frac{[A_3]}{[A_2]} = \frac{\sqrt{R_2^2 + \left(\omega_1 L_2 - \frac{1}{\omega_1 C_2}\right)^2}}{\sqrt{R_3^2 + \left(\omega_1 L_3 - \frac{1}{\omega_1 C_3}\right)^2}} = \frac{\omega_1 L_2 - \frac{1}{\omega_2 C_2}}{\omega_1 L_3 - \frac{1}{\omega_2 C_3}} \quad (22)$$

taking $R_2 = 0$ and $R_3 = 0$, and putting $C_1 L_1 = C_2 L_2$ and $\omega_1 = \omega_4$, from (20)

$$\frac{[A_3]}{[A_2]} = -\left(1 + \frac{L_2}{L_1}\right) \quad (23)$$

To fulfil the first condition and to improve the receiving efficiency, we have to make $\frac{|A_3|}{[A_2]}$ larger or $\frac{L_2}{L_1}$ as large as possible, that is, to make the transmitter inductance large compared with the antenna inductance. In the arrangement shown in Figure 2, the secondary coil L_2 of the coupling transformer has generally optimum value of self inductance required to obtain the maximum sending antenna current and the generated wave is one of the "coupling waves," as known from the principles of the pliotron oscillator.[11]

The treatment in this case, that is, when the coupling $L_2 L_2'$ is fairly close, will become more complex; but the same result as above for the best adjustment of the antenna system will be obtained by a similar treatment.

5. EFFECT OF THE BALANCING COIL

Notwithstanding the various adjustments which were tried to minimize the effect of induction on the receiving circuit, it was practically impossible to get rid of the induction entirely, and the balancing coil was found to be effective in cancelling out even small traces of induction. The induced electromotive force tends to ruin the detector, and the electromotive force of the balancing coil to balance it out must be the same in magnitude and opposite in phase.

From equation (8), the phase difference ϕ between A_2' and A_3' is:

$$\phi = tan^{-1}\frac{S_1'+S_3'}{R_1+R_3} + tan^{-1}\frac{S_1'}{R_1}$$

Thus, in the adjustment, $S'=0$, and we have

$$\phi = tan^{-1}\frac{S_3'}{R_1+R_3}$$

where S_3, generally speaking, is very large compared with R_1 and R_2, and therefore

$$\phi = \frac{\pi}{2} \text{ approximately.}$$

But it is nearly impossible to make $S'=0$ rigorously, and when S' is large compared with R_1, then

$$\phi = \pi \text{ approximately.}$$

It will be seen from this result that the counter-balancing coil is less effective in the former case than the latter, and its effectiveness depends upon the value of ϕ which can be varied from zero to π according to the adjustment. But in the arrangement shown in Figure 2 the writer found from test that the balancing coil works quite satisfactorily when it is coupled to the secondary coil of the receiving oscillation transformer.

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[11] T. Kikuchi, "Proc. Physico-Math. Soc. Japan," volume 1, page 14, February, 1919.

The disturbance in the receiver circuit consists of two currents, one from the branch current from the sending coil, and the other being the current induced in the receiving system directly by the transmitting antenna current. The balancing coil is quite effective in balancing out the latter even tho it may not eliminate the former. Generally speaking, the balancing coil is sometimes very effective, but it is not absolutely necessary when the circuit is in its best adjustment, especially when the receiving apparatus operates with regenerative action.

Consequently, to get the best adjustment for simultaneous sending and receiving, we choose the natural wave length of the antenna ($C_1 L_1$) equal to that of the transmitter circuit ($C_2 L_2$), and make the aerial capacity (C_1) as large as possible and at the same time the transmitter capacity (C_2) as small as possible.

6. MULTIPLEX SYSTEM OF WAVE TELEPHONY AND TELEGRAPHY

The above duplex system has proven to be of great utility, enabling (1) successful simultaneous radio telephone and telegraph communication between two radio stations, (2) successful simultaneous communication between a radio station (for example, on board ship, and a land wire telephone subscriber), radio and wire communication being automatically relayed in both directions (3), any number of multiplex wave telephone and telegraphs superposed on an ordinary telephone and telegraph line, and (4) any number of multiplex wave telephone and telegraph superposed on a power transmission line.

These various uses of wave telephony and telegraphy have been developed in Japan for the last three or four years.

As the result of many experiments, we are now (1) installing a station in Kobe to enable persons on ships to speak directly with the land wire telephone subscriber in Kobe, Osaka, and so on, (2) installing instruments in Tokyo, Yokohama, Osaka, and Kobe to provide a multiplex system of wave telephony and telegraphy by superposing modulated radio frequency waves on the telephone wires. These arrangements will be opened to the public before the end of March, 1919.

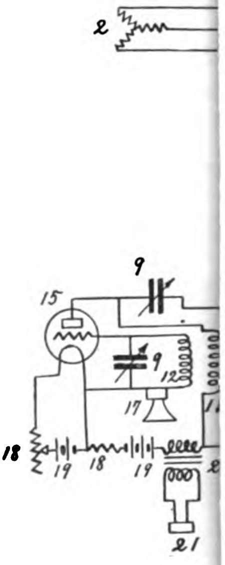
Generally speaking, telephone lines used near power transmission lines are subject to marked disturbances from induction, but most of this induction is of low frequency and causes no interference with wave telephony even when the wave is superposed on the power line itself for 50,000-volt transmission systems.

Dr. Torikata and his staff have carried out wave telephony over a power line successfully in May, 1918. The transmission power line of the Kinugawa Hydro-Electric Company, 90 miles (144 km.) in length was used. Since December, 1918, their system has been used in practice by the Fuji Hydro-Electric Company.

Details of the developments in certain directions are not yet available for publication, but the writer has been permitted to explain the operation of the guided wave telephone and telegraph, now successfully used by the Fuji Hydro-Electric Company, to the readers of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS.

7. "WIRED RADIO" TELEPHONE AND TELEGRAPH OVER POWER TRANSMISSION LINE

Figure 14 is the actual connection diagram of the wired wave telephone and telegraph used by the Fuji Hydro-Electric Company. In the Figure, (2) and (23) are the power stations connected to the 22,000-volt, 3-phase transmission line, (3) the receiving substation and (24) any intermediate station. Condenser (6) is designed to be quite safe at the line voltage, and inductance (7) is made nearly a dead short-circuit at the power frequency and a practically perfect choke coil at the radio telephone frequency. Oscillations are produced in the entire line by the bulb (15), which is also used as the receiving bulb, at station (2). Any two stations of (2), (3), (23), and (24) may speak with each other using the wave produced by the bulb(15). Calling may be accomplished by using the bulb amplifier and a loud speaking telephone, but may also be obtained by using the small transformer (26), the quenched spark gap (25), bulb rectifier (28), relay (29), and call bell (30). Arrangements are made to use sounder (36) and key (33) for wave telegraphy. At the intermediate stations, portable sets provided with "Koseki" or crystal detectors (34), microphone transmitters (17), and telephone receivers (21) are used. In the arrangement shown in Figure 14, the oscillations generated by the bulb (15) connected as in the Figure, are supplied to the overhead transmission line (1) thru the oscillation transformer (10). The amplitude of the current then flowing thru the line is modulated by the microphone transmitter (17) at the speaking station, and this enables the mutual or two-way conversation between any two stations, some of which may have a transmitting bulb, as at station (2), or which may not have such a bulb as at stations (3), (23), and (24). It is evident that conversation between two stations, both of which have transmitting bulbs, will give excellent results, but the sending arrangement in this case gives rise to much more expense and trouble. Consequently, the arrangement shown in Figure 14, which enables two stations, such as (23) and (3), neither having a transmitting bulb, to converse with each other, was adopted by Dr. Torikata for the Fuji Hydro-Electric Company. The working principle of this system is that, in guided radio



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Thus in guided radio signaling we can modulate the current in the entire line by varying the absorptions, resistances, or other constants of the whole system by a microphone transmitter, for example, and carry on two-way speech even without using the connections and careful adjustment indicated in Figure 2.

In conclusion, the writer wishes to express his hearty appreciation and thanks to Dr. Wichi Torikata for kind co-operation in the preparation of this paper.

SUMMARY: After reviewing the methods of duplex radio communication previously proposed, the author describes a divided-branch antenna arrangement with supplementary balancing coil, for this purpose. The theory and mode of adjustment of the system are explained.

The application of this method in practice to ship-to-shore radio telephony is described, and reference is made to the necessary wire-to-radio and reverse transfers.

Guided radio telephony and telegraphy over high tension transmission lines is considered. The use of a single oscillation generator for a number of stations is shown to be possible. An actual transmission of this type is described in detail.

RADIO FREQUENCY ALTERNATORS*

BY

MARIUS LATOUR

(This article is partly a reproduction of a communication recently presented before the Societé Française des Electriciens[1] by Marius Latour. It describes the different types of alternators capable of practical application to the direct production of radio-frequency currents. The author pays particular attention to the French designs of alternators, which he describes as "homopolar alternators with partial utilization of the periphery" and "homopolar variable reluctance alternators.")

In the last few years alternators of considerable output (exceeding 100 kilowatts) and of a frequency corresponding to the natural frequency of radio antennas have been built. The direct feeding of antennas from alternators without transformation of any kind and with the advantages inherent in sustained oscillations has thus become possible. Numerous papers on this subject have appeared. Mr. Bethenod among others has presented a particularly interesting paper before the Congress of Lyons in June, 1914.

This article intends to give a general survey of the different types of alternators that can be utilized practically for the generation of radio frequency currents.

The method of multiplying the frequency of the generator by means of transformers is not considered. It will suffice to remark that this method (used by the Telefunken Company) is based on the saturation of the magnetic circuit of a transformer by direct current for the purpose of producing harmonics in the curve of the primary alternating current. This method has been studied in France, as regards the second harmonic, by Leonard and Weber, and by Maurice Joly (who pointed out the importance of the method for the production of high frequency),

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* Received by the Editor, July 17, 1919.

[1] "Bulletin de la Societé Française des Electriciens," February, 1919, volume 9 (3d series), pages 97–115.

and, as regards the third harmonic, by J. Bethenod.[2] This method may form the subject of another article. The present paper considers exclusively alternators generating radio frequency current directly and without the help of supplementary apparatus.

The different systems capable of industrial application will here be classified as follows:

(1) Alternators in cascade.
(2) Internal Cascade Alternators.
(3) Homopolar disc alternators.
(4) Homopolar variable reluctance alternators.
(5) Alternators with partial utilization of periphery.

ALTERNATORS IN CASCADE

General Ferrié, as early as 1904, took an interest in the industrial production of radio frequency currents and has set up the problem of the generation of currents of the order of 80,000 cycles per second. Alternators giving 10,000 cycles per second were already known at that time; Lamme, for instance, has described before the American Institute of Electrical Engineers in 1904 a homopolar machine (to which I will come back later), which at a comparatively moderate peripheral speed gave several kilowatts at 10,000 cycles per second.

The generation of a frequency of the order of 80,000 cycles seemed to me to require new methods, and I proposed at that time the connection of a number of alternators in cascade. The arrangement is shown in Figure 1.

Let A_1-B_1, A_2-B_2, A_3-B_3, A_4-B_4 be four machines with alternate poles, mounted on the same shaft and having the same number of poles. Both the stationery and moving parts carry two-phase windings. A_1-B_1 being excited by direct current in A_1, gives two-phase currents of frequency of frequency f in B_1. These two-phase currents of frequency f are collected by means of rings not represented on the figure and made to feed the two-phase winding of the inductor A_2 of A_2-B_2 in such a way as to produce a revolving field in a direction opposite to that of rotation of the shaft: two-phase currents of frequency $2f$ are then generated in B_2. These latter currents are in their turn made to supply the inductor A_3 of A_3-B_3 in such a way that the revolving field created moves in a direction opposite to that of rotation of the rotation of the shaft: the current obtained in B_3

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[2] The Bethenod process differs from that of Joly in that the latter uses the alternating current itself to produce saturation. The Joly process is best adapted to the production of relatively low frequency.

has the frequency $3f$. These currents of frequency $3f$ are fed into A_4 of the machine A_4-B_4. The current of frequency $4f$ in armature B_4 is used to supply the antenna. By increasing the number of alternators, one can multiply the frequency indefinitely. With n alternators a frequency of nf is obtained, where f is the frequency of each single alternator.

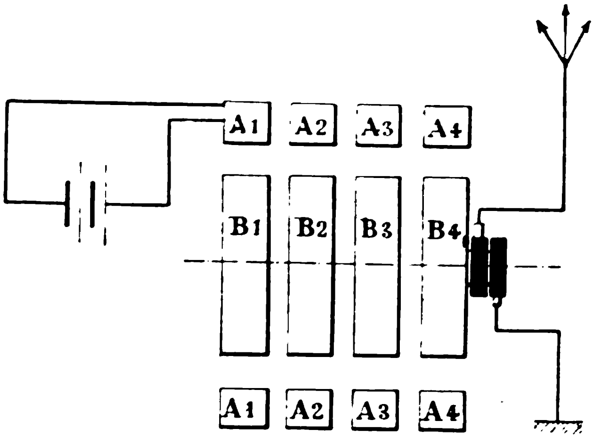

FIGURE 1

The use of collector rings between the successive circuits and at B_4 can be avoided by making the stationary and revolving parts alternately inductor and armature. In the figure A_1, B_2, A_3, B_4 would be made inductors; B_1, A_2, B_3, A_4—armatures. Between the different alternators condensers can be interposed to compensate for the reactance of the windings. When I first conceived the idea, different priorities were brought to my attention and I abandoned for the time being the project. The different new elements introduced by me, such as multiplication of frequency by means of polyphase instead of single phase windings and the avoidance of collector rings are, however, of real practical value. No attempt to carry out this idea in practice was made either in France or abroad until 1912. At that date Bethenod constructed the first high frequency alternator based on the principal of Figure 1. In this experimental alternator each elementary machine had a frequency of 6,000 cycles per second at a peripheral speed of 120 meters (393 ft.) per second. Altho the sheets used (silicon steel 0.25 mm. or 0.01 inch thick) were such as to cause high losses at the frequency used, Bethenod has obtained on the antenna a power of the order of one kilowatt at the frequency $4 \times 6000 = 24{,}000$ cycles per second. The system of Figure 1 operates as follows. The first alternator $A_1 - B_1$ acts only as a generator, the second alternator $A_2 - B_2$ acts one-half as a generator and one-half as a transformer; the third $A_3 - B_3$ one-third as a generator and two-thirds as a transformer, the fourth alternator one-fourth as a generator, and three-fourths as a transformer. Generally, the nth alternator acts to 1-nth part as a generator and to one-$\frac{n-1}{n}$th part as a transformer.

INTERNAL CASCADE ALTERNATORS

P. Boucherot has announced in 1893 (see "La Lumière Electrique" of the 25th of March, 1893, page 544) the following theorem:

"In an alternator with alternate poles delivering simple alternating current the armature is the seat of an electromotive force and a current represented by an infinite series of the odd terms of a Fourier series, the inductors are the seat of an electromotive force and a current represented by an infinite series of the even terms of a Fourier series."

Boucherot writes as follows:

"The coefficients of the currents of successive frequencies are naturally decreasing very rapidly on account of the self inductance of the two circuits, which reduces the intensities the more the higher the frequency. This is, however, not to be regarded as inevitable; it is possible to make the coefficients increase instead of decreasing; it is equally possible by utilizing the remarkable properties of condensers to cause these coefficients to increase up to the nth term and to decrease afterward. This allows the production of currents of very high frequency and the construction of alternators giving all sorts of frequencies but such that the currents of a given frequency are the most intense and can be considered as the main currents."

The Goldschmidt machine represented schematically by Figure 2 is a practical realization in 1907 of the ideas expressed by Boucherot in 1893.

The stator and rotor of the single phase radio frequency alternator of Goldschmidt type are indicated on the figure by the letters S and R.

In this alternator the *odd* frequencies f and $3f$ in the rotor and the even frequencies $2f$ and $4f$ in the stator are reinforced; this latter frequency $4f$ becomes, properly speaking, the utilized frequency. The rotor circuit is closed over such a system of inductance and capacity that it can simultaneously enter into resonance with the two odd frequencies f and $3f$. To accomplish this, the shunt L-C is so proportioned as to form a short-circuit for the frequency f; the capacity C_1 is of such value as to balance the reactance of the rotor at the frequency f, and the capacity C_2 is such that, combined with the shunt L-C and the capacity C_1, it balances the reactance of the rotor at frequency $3f$. The stator is excited by means of direct current thru the protective reactance λ, and is connected to a system of inductances and capacities L', C'', C_1', C_2' which allows resonance with the frequencies $2f$ and $4f$. The capacity C_2' is in practice represented by the antenna circuit.

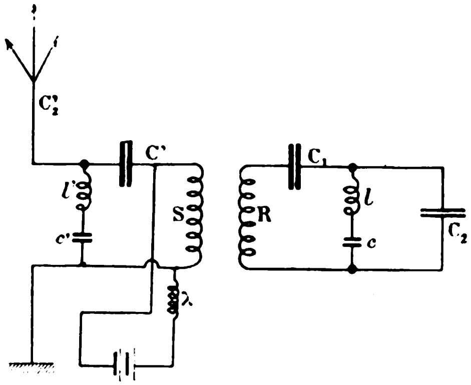

FIGURE 2

This Goldschmidt alternator seems to have exerted a fascination on the investigators of all countries, and is often spoken of as the "remarkable" Goldschmidt machine. This alternator does not, however, differ either in principle or in practical value from the system of alternators in cascade.

The current of frequency f generated in the rotor produces an alternating field, which, by Fresnel's theorem, can be decomposed into two revolving fields of half magnitude. One of these fields rotates in the same direction as the rotor and therefore gives rise in the stator to a current of frequency $2f$ in exact accordance with the principle of alternators in cascade. This current of frequency $2f$ sets up an alternating field which in its turn is decomposable into two revolving fields of which one is rotating in a direction opposite to that of the shaft, and again in accordance with the cascade principle, generates a current of frequency $3f$ in the rotor. This current of frequency $3f$ in its turn produces a revolving field of frequency $3f$, which, taken along by the rotor, gives rise in the stator to a current of frequency $4f$. To sum up, the alternator of Figure 2 is essentially an assembly in one machine of the four machines mounted in cascade of Figure 1. The concentration of several machines in one is however irrelevant both from the point of view of efficiency and of specific power as we had already occasion to prove elsewhere (see "La Lumière Electrique" of June 22, 1912, page 357).

Let us first consider the iron losses, and let us designate by p the losses caused by a field ϕ rotating in a positive or negative direction with the velocity ω—with respect to the iron. Fields $\phi_1, \phi_2, \phi_3 \cdots$,—rotating with velocities $\omega_1, \omega_2, \omega_3 \cdots$,— differing in magnitude and direction, will then, if considered separately, cause losses p_1, p_2, $p_3 \cdots$ If all these revolving fields coexist, the total loss is simply equal to the sum: $p_1+p_2+p_3 \cdots$

Likewise, for the copper losses, let us designate by ω the Joulean losses in the windings due to the system of currents i producing the field which rotates with the velocity ω in a positive or negative direction. Systems of currents i_1, i_2, $i_3 \cdots$ giving rise to fields revolving with velocities differing in magnitude or direction will cause, if considered separately, the losses ω_1, ω_2, $\omega_3 \ldots$ If all these different current systems coexist in the same windings, they cause total losses which are simply equal to the sum of the separate losses $\omega_1+\omega_2+\omega_3+ \cdot$

When the losses of the separate machines are combined in a single machine, it is necessary in order to limit the temperature rise to the same value, to give to the single machine a size equal to that of the several machines placed side by side and no beneficial effect is secured.

From these considerations, that I have formulated in "La Lumière Electrique," as far back as 1912, it follows that in radio frequency alternators with alternate poles as well as in all alternators operating far from saturation, the electrical efficiency is independent of the heating effect, that is, of the overload. If the inductor flux and the output are doubled, the losses are quadrupled, but so also is the power. The ratio between the losses and the power remains constant. The power per unit volume is thus limited by the facility with which the generated heat can be dissipated. The advantage of artificial cooling becomes apparent. This cooling is the more advantageous as the great peripheral speeds used in radio frequency alternators cause large ventilation losses which are relatively the less the larger the power output.

In order to diminish the ventilation losses I have proposed (see "La Lumière Electrique," June 22, 1912, page 356) to run the alternator under reduced air pressure. It is for experiment and test to weigh and overcome the complications involved in this proposal.

The advantage of the Goldschmidt machine is that it permits the use of a single-phase winding on the rotor, its disadvantage is that it does not permit of subdivision of power as does the external cascade system.

HOMOPOLAR DISC ALTERNATORS

The homopolar alternator has no windings on the moving parts. This is of great advantage at high peripheral speeds, and this type of alternator was therefore used from the very first for the generation of radio frequencies. Thury has built between 1893 and 1900 about a dozen homopolar alternators giving currents of 10,000 cycles per second. One of these alternators was exhibited at the Exposition of Geneva in 1896 (see "L'Eclairage Electrique" of October 24, 1896, page 157). This alternator was of the "bell" type favored by Thury at that time. The revolving inductor carried 200 polar projections and run at 3,000 revolutions per minute for a frequency of 10,000 cycles per second. It delivered 3 to 4 kilowatts, at this frequency and had an open circuit voltage of 150 volts. Thury used both solid and laminated inductors.

In 1904, Lamme described in detail in the "Transactions of the American Institute of Electrical Engineers" a homopolar alternator built by him. This alternator is of the normal type, but in its construction are used for the first time specially thin steel sheets of only 0.075 mm. (0.003 inch) in thickness. The diameter of the inductor was 62 cm. (24.4 inches), the number of revolutions 3,000 per minute; the peripheral speed was thus 100 m. (305 feet) per second. It would seem that by pushing further the excitation of the machine and by compensating the reactance of the armature by a series condenser, a much larger power output than that given by Lamme (2 kilowatts) could be obtained.

In order to utilize higher peripheral speeds up to 200 and 300 meters (610 and 915 feet) per second, Alexanderson has designed the disc type of the homopolar alternator. Figure 3 represents an Alexanderson homopolar alternator. The winding *B* creates a continuous field which traverses the disc *A* and the laminated armature *C*. The disc carries on either side radial polar projections; the armature carries a winding, the conductors of which are placed in radial slots.

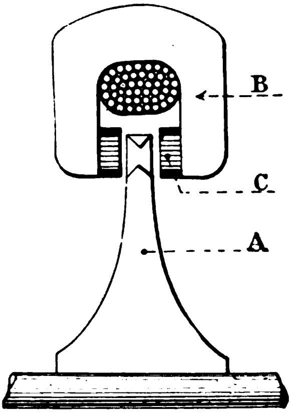

FIGURE 3

Alexanderson first sought to reduce the ventilation loss by inserting non-magnetic pieces between the polar projections of the disc so as to obtain a perfectly smooth surface. Later on he replaced these non-magnetic pieces by conducting pieces connected externally in such a way as to form a sort of a squirrel-cage on each face of the disc. Such an amortisseur system is capable of reducing the losses due to the machine output.

One could, as a matter of fact, place the laminated polar pieces in openings secured in a solid bronze disc as shown in Figure 4, which construction would reduce the losses in the moving part to a minimum.

The open-circuit characteristic of homopolar machines has an interesting particularity that has been pointed out long ago (for instance by Alexanderson, "Proceedings of the American Institute of Electrical Engineers," 1909, page 410). While the characteristic of alternators with alternate poles is somewhat similar to that of the magnetisation curve of iron, that of the homopolar machines has the form represented on Figure 5. The open-circuit voltage first increases with the excitation, passed thru a maximum and then diminishes.

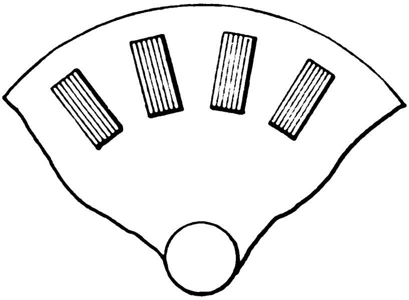

FIGURE 4

The reason of this particular behaviour lies in the fact that saturation acts in two different ways; first in the usual way, by diminishing the ratio of the flux-increase to the excitation increase, and second, by so to say effacing the poles of the rotor, since the interspaces finally begin to emit flux as well as the polar projections. The difference of flux emission between the vacant spaces and the poles which constitutes the flux variation in the alternator tends thus to disappear as saturation is approached.

As a result of these two causes there exists a voltage maximum for a certain excitation. To get the maximum power out of the alternator it is necessary to operate near this maximum voltage, which does not correspond in general to an excessively large variation of flux in the armature teeth. Under these conditions it is not justifiable to say that the electric efficiency of the homopolar alternator is independent of the load. This was true of the machines with alternate poles, but the simple reasoning holding in that case is no more applicable. The losses are no longer proportional to the output, and consequently the electric efficiency diminishes with overload.

Another remark regarding the hysteresis losses in homopolar machines is here appropriate.

According to measurements by John D. Ball, reported by Steinmetz, ("Theory and Calculation of Electric Circuits," 1917, page 76), the hysteresis losses on passing from induction B_1 to induction B_2—corresponding to an alternating variation of induction $\frac{B_1-B_2}{2}$—are the greater, the larger the mean induction $\frac{B_1+B_2}{2}$. In other words, the Steinmetz coefficient depends on the mean value of induction around which the flux variation takes place. Figure 6 gives η as a function of $\frac{B_1+B_2}{2}$. From this figure we may derive, for instance, that at the mean induction of 10,000 gauss the hysteresis losses are multiplied by 2.3, at 14,000 gauss by 3.5.

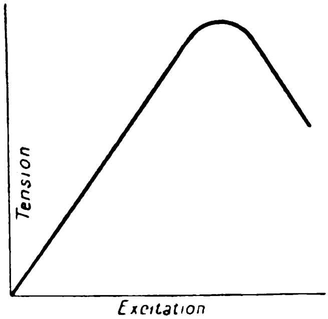

FIGURE 5

We conclude from this that the hysteresis losses are larger in homopolar alternators than in alternators with alternate poles, and that therefore the subdivision of the iron or changes in its silicon content, which reduce only the eddy losses, are of less importance.

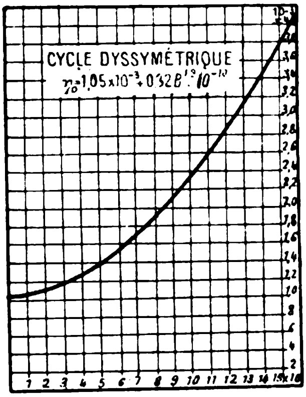

FIGURE 6
Asymmetric Cycle, $\eta_o = 1.05\ (10)^{-3} + 0.32\ B^{1.9}\ (10)^{-10}$

HOMOPOLAR ALTERNATORS WITH VARIABLE RELUCTANCE

One form of this type of alternator is shown in Figure 7. It comprises an excitation winding on the stator, which, as pointed out by Bethenod, may at the same time constitute the armature winding, provided an inductance is inserted as shown in the figure. When the rotor teeth are facing the stator teeth there is maximum flux in the armature; when the interspaces are facing the teeth, the flux is at its minimum. This produces an electromotive force of the same frequency as that of a homopolar alternator having the same number of teeth or polar projections on the rotor, and only half the number of slots on the stator. The slots must, however, in this case allow the passage of direct current. The rotor is necessarily laminated as it is subject to a varying flux.

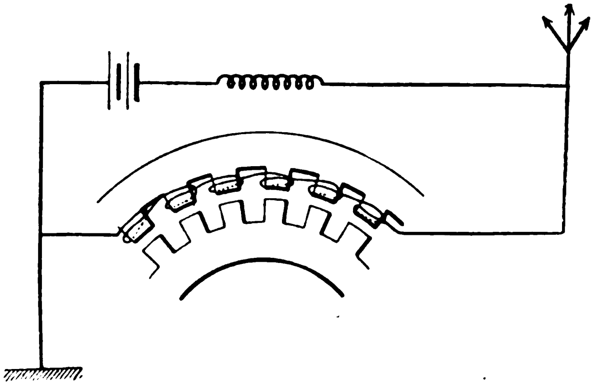

FIGURE 7

To obtain high frequency the rotor teeth must be very thin and closely spaced and the flux variation effective in the armature becomes very small in comparison with the total flux. The flux changes its sign in each rotor tooth for each displacement corresponding to the pole pitch and as a result the rotor losses may become much larger than the stator losses. The efficiency of this form of variable impedance alternator soon becomes inferior to that of the homopolar alternator, the polar of which projections are always traversed by a uni-directional flux.

The variable impedance alternator can, however, also be given the homopolar form.[3] It is sufficient to imagine a homopolar alternator entirely laminated in a direction parallel to the shaft, having a stator with open slots without windings carrying a number of teeth equal to that on the rotor (see Figures 8 and 9). The central excitation coil is then traversed exactly as in Figure 7 by a maximum or minimum flux according to the position of the rotor teeth in relation to the stator teeth. The flux does not now change its direction in the rotor teeth and the advantages of the ordinary homopolar alternator as regards losses are present.

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[3] See Latour, British patent number 102,738.

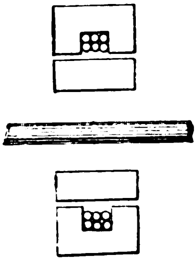

FIGURE 8

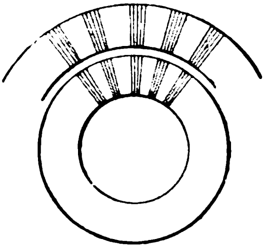

FIGURE 9

Theoretically, as there are no conductors to be placed in the vacant spaces of the stator, this alternator could be built for any desired frequency. For instance, with a peripheral speed of 150 meters (458 feet) second and teeth of the order of one millimeter (0.04 inch), a frequency of 100,000 cycles per second would be reached. In reality one would have to make the mechanically impossible assumption that the air-gap diminishes with the width of the teeth. It is easily seen that as soon as the width of a tooth becomes of the same order of magnitude as that of the air-gap, the reluctances for the two extreme positions of Figures 10 and 11 become about equal. With an air-gap of 0.5 mm. (0.02 inch) it is feasible to have teeth 2 mm. (0.08 inch) wide and to obtain directly a frequency of the order of 40,000 cycles per second. At the present time, I have under construction homopolar generators of variable reluctance, giving at a peripheral speed of 150 meters (457 feet) per second, a frequency of 60,000 cycles. These are covered by British patent number 102,738; and the United States patent is expected to appear soon.

It is possible to increase the difference between the maximum and the minimum flux correspondingly to the two extreme positions of Figures 10 and 11, by disposing in the polar interspaces of the stator and of the rotor conducting elements which serve as amortisseurs to the flux emitted by the teeth flanks.

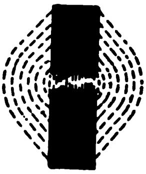

FIGURE 10

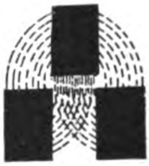

FIGURE 11

In the position of Figure 10 (maximum flux) each tooth emits one flux in the air-gap and two identical lateral fluxes. In the position of Figure 11 (minimum flux) each tooth emits four identical fluxes which link respectively a tooth-face with a tooth-flank.

Let us consider the excitation produced by unit current and let us call

Φ—the air-gap flux of a tooth in the position of Figure 10.

ϕ—the lateral flux emitted by each tooth-flank in the position of Figure 10.

ψ—the flux linking a tooth-face with a tooth-flank in case of Figure 11.

The flux enclosed by the armature varies between the values $(\Phi+2\phi)$ and 4ψ.

With conducting elements acting as dampers, the flux emitted by the flanks of the teeth cannot vary; as a result of the damping currents this flux will remain fixed at the average value of $\frac{\phi+\psi}{2}$.

The flux will therefore oscillate between the values

$$(\Phi+\phi+\psi) \text{ and } 2(\phi+\psi)$$

If, as a first approximation, one assumes, as in the figures, $\psi=2\phi$, one finds that without amortisseurs, the flux oscillates between $(\Phi+2\phi)$ and 8ϕ, and with amortisseurs, between $(\Phi+3\phi)$ and 6ϕ.

In the first case the flux variation is $(\Phi-6\phi)$, in the second $\Phi-3\phi$.

By a numerical example one can readily form an estimate of the very considerable improvement produced by the introduction of amortisseurs.

By reducing the reaction flux and the losses caused thereby the amortisseurs also have an important influence on the output $I \sin \omega t$. It is easily seen, that the only alternating reaction flux possible is the one corresponding to the air-gap flux Φ proper. The flux emitted by the flanks of the teeth is of the form

$$I \frac{\psi+\phi}{2} \sin \omega t + I \frac{\psi-\phi}{4} [\cos a - \cos (2 \omega t - a)]$$

and on account of its alternating character is reduced to zero by the amortisseurs. The continuous part of the flux causes no losses.

The variable reluctance alternator is able to multiply the frequency in a way analogous to the Goldschmidt alternator. The basis of this is formed by the second theorem of Boucherot ("La Lumière Electrique" of March 4, 1893, page 500), which is as follows: "In an alternator with non-alternate poles the inductor and the armature carry currents represented by all the even and odd terms of the Fourier series."

The variable reluctance alternator can thus give the highest frequencies without excessive peripheral speeds.

For instance, considering the doubling of the frequency, the output of current $I \sin \omega t$ creates a reaction flux

$$\frac{I \Phi}{2} \sin \omega t - \frac{I \Phi}{4} [\cos a - \cos (2 \omega t - a)]$$

which causes the appearance of a voltage of double frequency

$$-\frac{\omega I \Phi}{2} \sin (2 \omega t - a)$$

In a variable reluctance alternator it is possible to work sufficiently far from saturation so as to have the efficiency independent of overload in the same way as in cascaded alternators. If, by placing an appropriate capacity in series with the armature, one brings the reaction flux and the voltage-generating flux (v) in quadrature, the electric losses on open circuit and those due to the load I will simply superimpose, and the best efficiency will obviously be obtained when the load-losses are equal to the open-circuit losses. This follows from the consideration that the available power is $v I$, while the losses[4] have the form

$$k' v^2 + k'' I^2.$$

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[4] It is well to note that the use of a series capacity to compensate for the total inductance of the armature does not lead in all cases to optimum efficiency. In order to work with a minimum resultant flux, and therefore with minimum iron losses for a given power, it is necessary to increase the excitation with the load and compensate with the capacity only the inductance due to leakage. The use of capacity for the compensation of inductance results in a reduction of copper losses and in certain cases, near saturation, is all that makes it possible to develop power. The above considerations apply to all radio frequency alternators.

ALTERNATORS WITH PARTIAL UTILIZATION OF PERIPHERY

The difficulty in designing ordinary homopolar or alternate pole alternators for radio frequencies without undue increase of peripheral speed is due to the practical impossibility of placing coils in the slots which soon become too narrow. For instance, with a peripheral speed of 150 meters (458 feet) per sec. and for 30,000 cycles per second, the pole-pitch is 2.5 mm. (0.1 inch). With a tooth-wide of only 2 mm. (0.08 inch) there is left only 0.5 mm. (0.01 inch) for the slot. To overcome this, the idea suggests itself to place in one alternator only a fraction of the needed poles and to combine several alternators among which the totality of poles is distributed. Instead of one alternator one could, for example, consider a system of three alternators I, II, III (see Figure 12), in which the successive poles are found by passing from one alternator to the next and two of every three poles are omitted in each alternator. The space left by the missing poles in each alternator then permits placing the coils more easily. Figure 12 represents, to the left, the rotors and, to the right, the stators of the three alternators, and is sufficiently explicit. A first north pole being in alternator I, the corresponding south pole is in alternator II, the second north pole in alternator III, the second south pole is again in alternator I, and so on. In each rotor, two out of every three poles are omitted. In the three corresponding stators two teeth out of every three are likewise omitted.

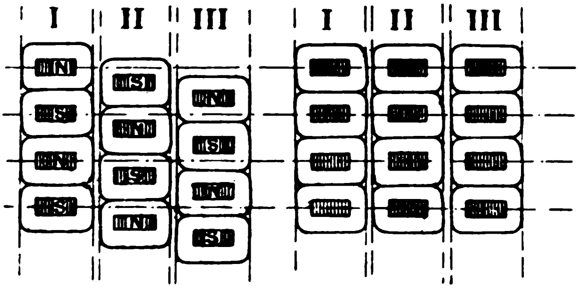

FIGURE 12

This construction is of general application and can be used in homopolar alternators and in non-homopolar variable impedance alternators such as that of Figure 7.

In these latter, there is no excitation winding on the rotor and the vacant spaces are to be considered as simply replacing the poles of opposite sign.

Figure 13 relates to homopolar alternators, Figure 14 to variable reluctance alternators.

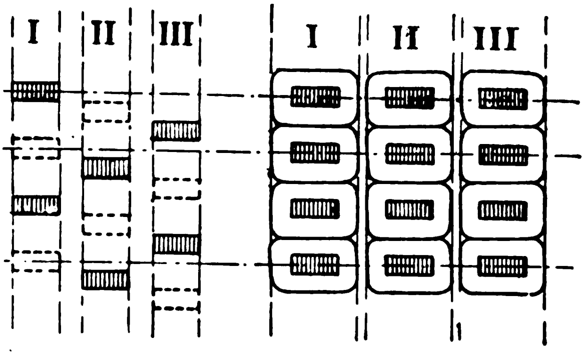

Figure 13

In thus distributing the poles of one alternator among three, it is assumed, as a first approximation, that the air gap is infinitely small, so that the flux emitted by each inductor pole is concentrated on its width facing the air gap, and that the flux issued from the flanks can be neglected. In reality, this latter flux may be quite appreciable, and therefore, beside the frequency which results from the combination of the three alternators, there will be a fundamental frequency given by each alternator which is ⅓ of the frequency given by the combination. In other words, the arrangement of narrow poles represented on Figures 12, 13, and 14 gives rise in each alternator to a third harmonic, which harmonic is brought out by connecting in series the three alternators displaced in phase $\frac{2\pi}{3}$. The idea of purposely producing the harmonics of alternators in order to obtain high frequencies is old, and dates back to Max Wien (1902). The explanations given above have permitted us to determine the width of the poles necessary to produce the third harmonic.

Beside the losses due to the harmonic frequency, there are in each alternator losses due to the fundamental frequency, which may be much more serious than the first mentioned.

The partial utilization of periphery may, however, be limited to the stator only, by sliding the three alternators of Figure 12 into one, in which case the fundamental frequency disappears. For the homopolar and the variable reluctance alternator, the arrangements shown on Figures 15 and 16 are respectively arrived at. In these alternators the polarity of the rotor is three times that of the stator. The frequency is determined by the polarity of the rotor[5].

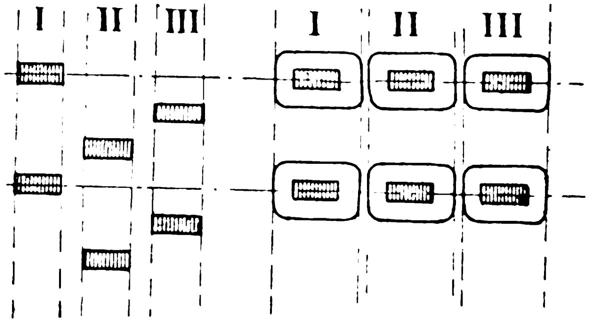

FIGURE 14

Figures 17 and 18 represent the contours of the stator and rotor of the homopolar and of the non-homopolar variable impedance alternator.

Bethenod and his assistant Billieux have built homopolar alternators on the above principle. The rotor was laminated, and constructed according to a design previously intended for

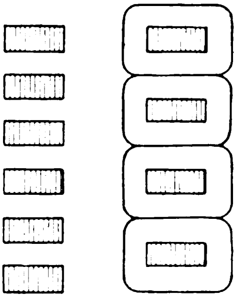

FIGURE 15

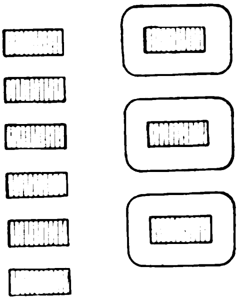

FIGURE 16

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[5] See my U. S. Patent 1,234,912.

the variable reluctance alternator, as in Figure 7 which was first suggested by Bethenod.

The work on the radio frequency alternators has been conducted in France during the war at the request of Colonel (now General) Ferrié, who ever since 1904 has interested himself in this subject.

All the possible solutions have been taken up in order, and we feel that, thanks to the instigation of the Military Telegraph Department, this question has been very thoroughly and practically conclusively investigated in this country.

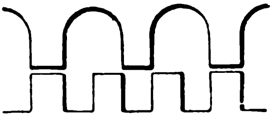

FIGURE 17

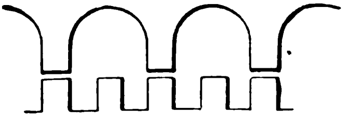

FIGURE 18

In Figure 19 is shown a 225-kilowatt, 20,000-cycle radio frequency alternator. This is a homopolar generator with partial utilisation of the periphery.

SUMMARY: The various types of radio frequency alternators are described. They are historically considered; and the general principles underlying their design are given. The special types most used in France at present are considered in detail.

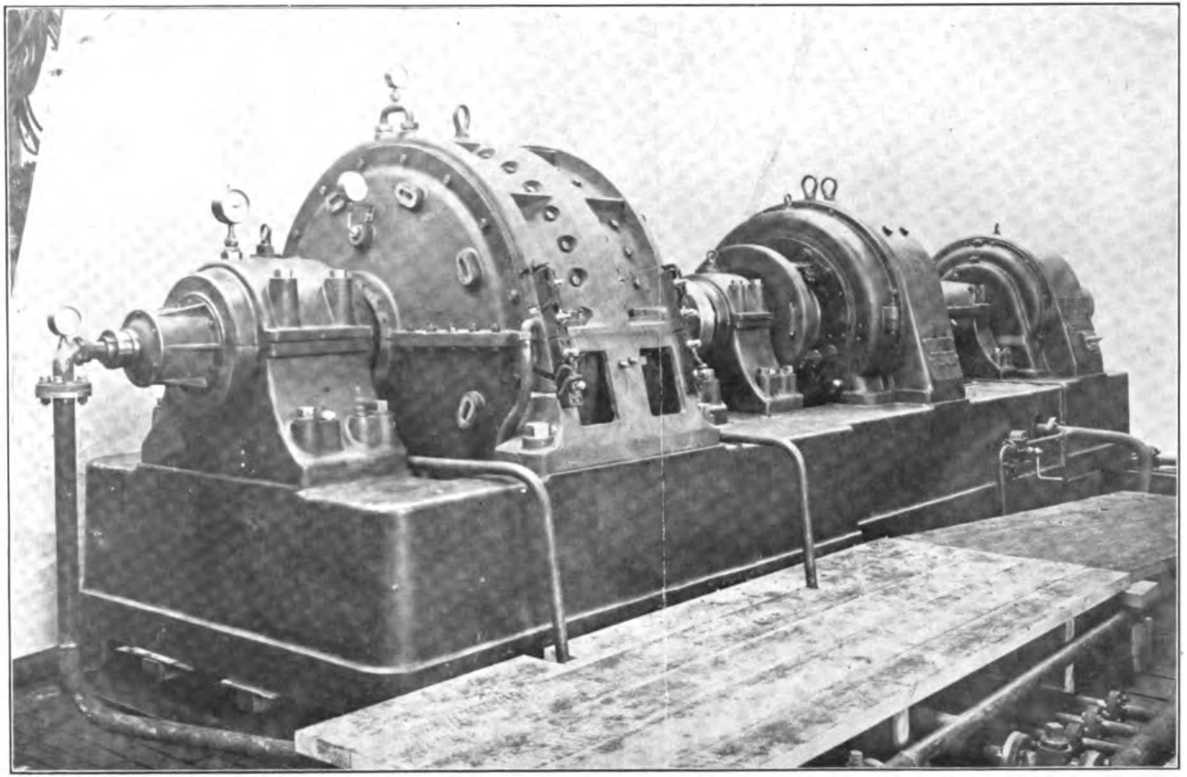

FIGURE 19—225-Kilowatt, 20,000-Cycle Latour Radio Frequency Alternator

SOME NOTES ON VACUUM TUBES*

By

JOHN H. MORECROFT

(ASSOCIATE PROFESSOR OF ELECTRICAL ENGINEERING, COLUMBIA UNIVERSITY, NEW YORK)

FORM OF PLATE CURRENT CURVE

The general theory of vacuum tubes has been well covered by Langmuir, van der Bijl, and others; it is the purpose of these notes to point out some of the more detailed actions of a tube with the idea of helping those who attempt to teach the theory and operation of tubes, and who will, unless they have tested many tubes, have difficulty in reconciling their experimental results with the theory as it has been presented to date.

The general expression for the plate current of a tube is

$$I_p = a\,(E_p + \mu_o\, E_g + \varepsilon)^x.$$

The quantity ε is very small, especially with oxide-coated filaments, as may easily be proved by connecting both the grid and plate to the negative end of the filament (that is, having both E_p and E_g equal to zero), and noting how much current flows to the plate. This is so small that a sensitive galvanometer is required to measure it with the ordinary tube.

One of the factors determining the value of ε is the velocity with which the electrons are expelled from the filament, which, increasing with temperature, is evidently greater for tungsten filaments than for oxide filaments. To show how the temperature affects this quantity, ε, a tungsten filament tube, having two electrodes only, had its plate connected to the negative end of its filament thru a microammeter, and the current thru the meter was noted for various filament currents. As there was no battery in the plate circuit, and it was connected to the negative end of the filament, the only force causing the electrons to flow from the filament to the plate was due to the velocity with which they were ejected from the hot filament. The result of this test is shown in Figure 1; the normal current thru the filament is 1.1 amperes. The plate current for this filament current

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* Received by the Editor, June 10, 1919.

[illegible] rapidly as [illegible]ere were more [illegible]elocity. It may [illegible]ity ε has, when it is [illegible]he plate current in[illegible]croamperes by merely [illegible] of the filament, thus [illegible]ament by perhaps 1 volt,

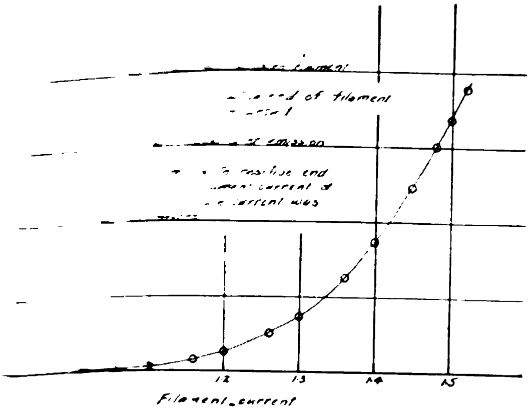

FIGURE 1

[illegible] disposed of ε as being generally negligible let us [illegible] that E_g is made zero by connecting the grid to [illegible] end of the filament, thus reducing the expression [illegible] current to the simple form

$$I_p = a\,E_p^{\,x}$$

[illegible] value of this exponent x has been given as 1.5 and 2.0, [illegible] a theoretically derived value and the latter an ex[illegible]ally determined value. Now for a certain class of [illegible] which will probably be extensively used in laboratory work, [illegible] those designed for low plate voltage, the exponent x [illegible] widely from both of the values given above. Figure 2 [illegible] the variation of plate current in an oxide-coated filament [illegible] designed for a plate voltage between 20 and 40, as the plate [illegible]tage was varied, the grid voltage being zero. On the same [illegible] is shown the value of the exponent x for various plate [illegible] It is seen that not only does the value not lie between

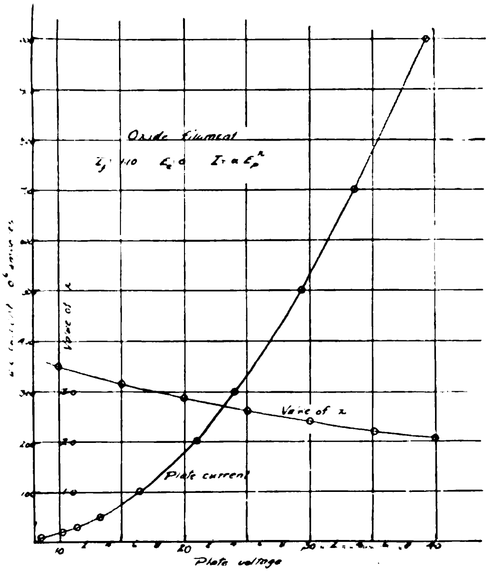

FIGURE 2

the values 1.5 and 2.0 but it varies with the plate voltage. Figure 3 shows the same condition for a tungsten filament tube designed for the same plate voltage.

In Figure 4 are shown these two curves plotted on logarithmic co-ordinate paper, as is also the straight line given by a curve having an exponent equal to 2.

Twelve tubes of each of the types used for the results shown in Figures 2 and 3 were selected at random from a large collection and the variation of plate current with plate voltage determined, and the results plotted on logarithmic paper; the result is shown in Figure 5. The twelve curves for each type lay inside the area indicated by the cross hatching, in general not lying parallel to the boundary lines of the areas, but crossing each other in haphazard fashion. It has of course been recognized that when the plate voltage becomes equal to, or less than, the *IR* drop in the filament, the ordinary form of equation does not hold, but the curves obtained from these twenty-four tubes differ in their exponents by a ratio of five to one, with plate voltages still greater than the filament *IR* drop.

One of the principal reasons for the variation of the exponent as the plate voltage varies is the fact that there is a different voltage between each part of the filament and the plate. This

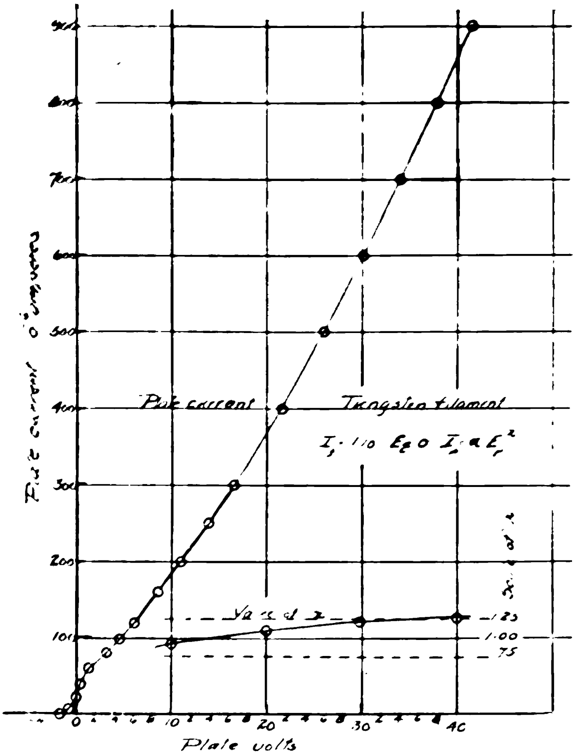

FIGURE 3

large change in the exponent does not occur if the surface emitting the electrons is an equipotential surface; such a tube, having an electrically heated equipotential surface was described and used by the writer some years ago.* With such a tube, the theoretical relations between space charge and plate voltage are more closely obeyed.

RESISTANCE OF A TUBE

The output resistance of a tube is the ratio of the change in plate voltage to the corresponding change in plate current, other conditions remaining the same. It is the reciprocal of the partial derivative of the plate current with respect to the plate voltage. The resistance of this circuit would ordinarily be obtained by dividing the plate voltage by the plate current as for any other circuit. This value of resistance however would be

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* "Physical Review," volume 8, number 5, page 563.

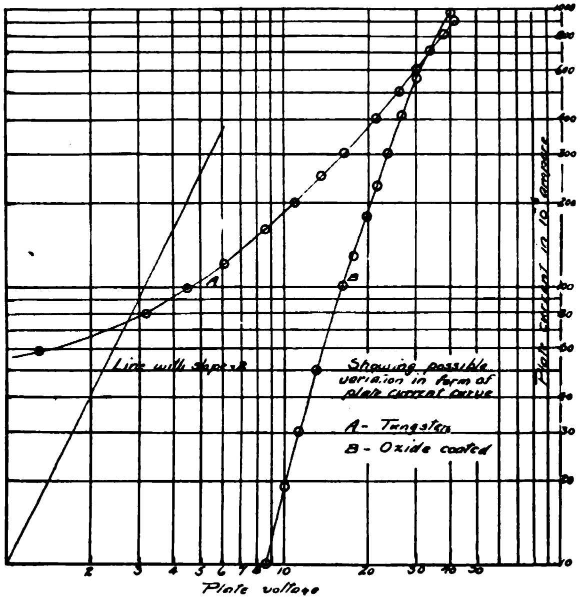

FIGURE 4

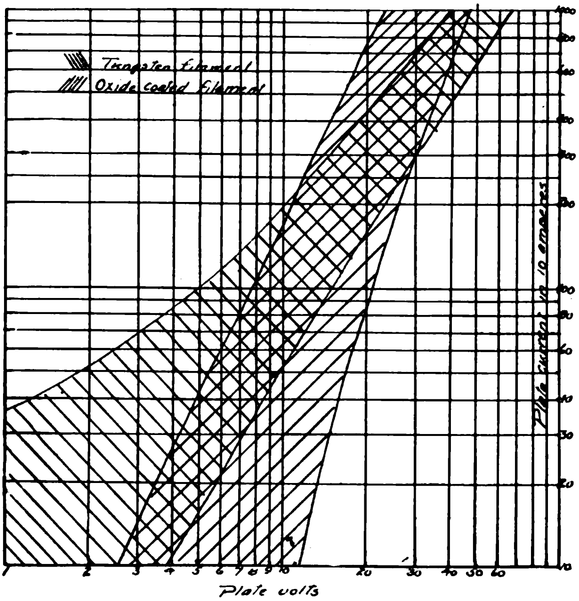

FIGURE 5

the resistance for continuous current only and is generally greatly in excess of the alternating current resistance, obtained as indicated above. However there is a simple relation between the continuous current resistance and the alternating current resistance. Calling the two values of resistance R_{ac} and R_{cc} it is seen that if $I_p = {}^{a}E_p{}^{x}$, then $R_{ac} = R_{cc}/x$. Hence it should be possible to calculate the a.c. resistance from the c.c. resistance, it being supposed that x is known.

In Figure 6 are shown the two resistance curves for a tube, the one being the quotient of the plate voltage and the plate current and the other having been obtained by the method first described by Miller, using a low voltage when making the measurement. Points indicated by circles are the experimentally determined points; on the curve of alternating current resistance are shown a series of points indicated by crosses; these are the values obtained by dividing the c.c. resistance by the value of x of the tube for that particular voltage, this x having previously been determined.

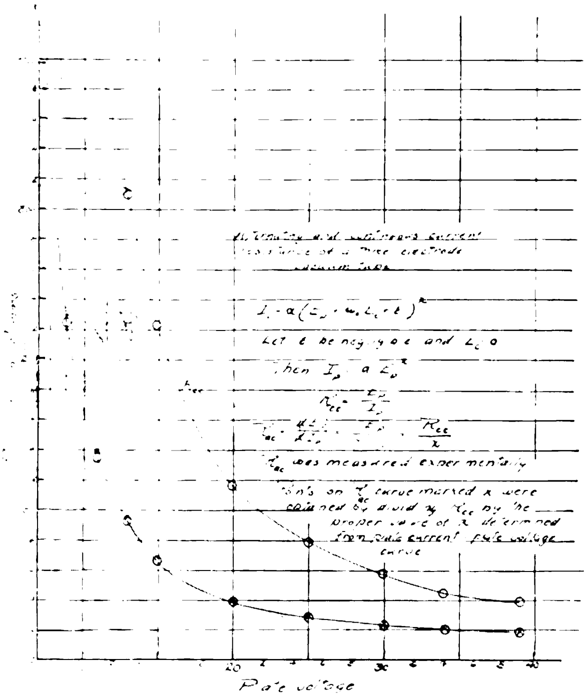

FIGURE 6

The value of R_{ac} will be constant thruout a certain variation of the excitation voltage. For large values of alternating voltage impressed on the grid, R_{ac} will increase and may become equal to R_{cc}.

DISTORTION EFFECT IN AN AMPLIFIER TUBE

If the grid of an amplifying tube is excited by an alternating emf. $E_g \sin \omega t$, and the plate current is of the form assumed by van der Bijl then the plate current becomes

$$I_p = a(\tau E_p + E_c + \varepsilon)^2 + 2a(\tau E_p + E_c + \varepsilon) E_g \sin \omega t + \frac{a E_g^2}{2}(\cos 2\omega t + \pi) + \frac{a E_g^2}{2}$$

as was pointed out by him. A direct current meter in the plate circuit will respond to the first and fourth terms only. The third term represents distortion and, as has been pointed out, it may be made as small as desired by adding sufficient resistance in the plate circuit. Now the third term has the same coefficient as the fourth, and as the fourth term is shown by a direct current meter it is possible to predict the distorting effect of a tube by making a test with a continuous current meter in the plate circuit. If, as the alternating emf. impressed on the grid is increased, the reading of the continuous current ammeter in the plate circuit remains constant, the third term of the plate current expression (that is, distortion) is absent.

Figure 7 shows the effect on the reading of a c.c. ammeter in the plate circuit of an amplifying tube as the alternating emf impressed on the grid is increased by small steps. Curve 1 shows the effect on the plate current when there is no added resistance in the plate circuit. The increase in the current is due to the fourth term of the current expression; this increase plotted on logarithmic paper with the grid voltage (E_g) as the other ordinate gives very nearly a straight line, the slope of which is 1.9. It is evident, therefore that the exponent 2 in the expression for plate current is nearly correct for this special tube within the narrow range of grid voltage used. Curves 2, 3, and 4 show the effect of adding resistance in the plate circuit keeping the voltage of the plate circuit, E_B, constant. The plate current is of course diminished as the plate circuit resistance is increased because of the decreased plate voltage brought about by the *IR* drop in the plate circuit resistance. Curve 4 shows a constant plate current as the grid voltage is varied thru a range of two volts.

Figure 8 shows how the plate current of a tube may be ex-

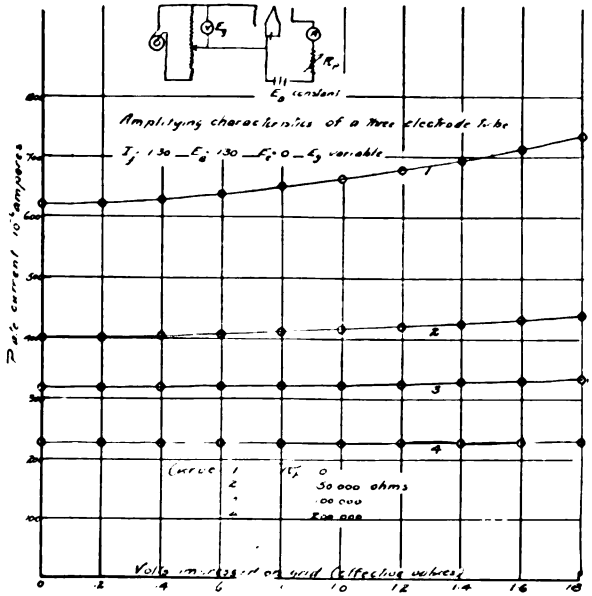

FIGURE 7

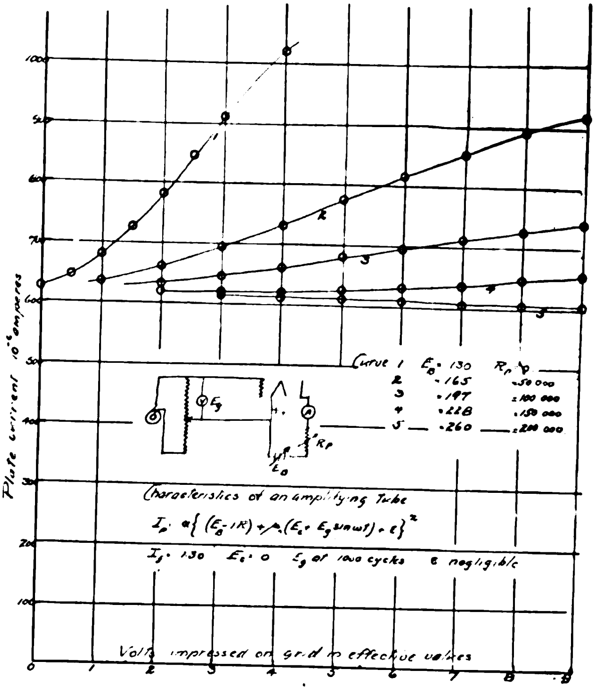

FIGURE 8

pected to vary (with an alternating voltage impressed on the grid) if various resistances are inserted in the plate circuit and the plate circuit voltage is sufficiently increased each time to give the same voltage on the plate, the grid voltage being zero. It is seen that if too much resistance is added in the plate circuit the plate current actually decreases as the grid voltage increases, thus again bringing about distortion.

CAPACITY AND CONDUCTANCE OF VACUUM TUBES

The internal capacity of a vacuum tube is of little importance at telephone frequencies, or at radio frequencies when large capacities are used in the tuning circuits of the tubes; but when the circuit condensers are small and the frequencies are high, the internal capacity may become of extreme importance. In the design of amplifiers and cetectors, this internal capacity plays an important role. In many laboratory experiments it had seemed to the writer that the capacity of the tube must be playing a much more important part than might be supposed, so a series of tests were undertaken to determine just how much the internal capacity of a tube might be and what conditions affected it. The work was started two years ago but it was stopped by pressure of war work on another problem.

As the capacities to be measured are of the order of a few micro-microfarads it is evident that some sort of differential method must be used to obtain accurate results; also to make the reactances to be measured of a reasonably small value radio frequencies must be employed. I tried at first to make the measurements by noting the change in the beat note of two small continuous wave sets with and without the capacity to be measured being connected in the circuit. I listened to the beat note without the desired capacity connected and then connected it in parallel with the condenser of the generating set and the received note changed. The note could then be brought to its original value by changing the condenser of the transmitter and the amount of this change noted. The scheme proved too difficult to manipulate, however; control of the voltage impressed on the tube to be tested and the separation of the effects of the conductance and the capacity made the method cumbersome and unreliable.

I therefore decided to carry out the measurements on the Wheatstone bridge, by which method the effects of conductance and capacity could be analyzed separately. Professor Pupin and his assistant, Mr. J. G. Aceves, had done a great deal of development work on the high frequency bridge, and it had proved to be reliable when proper precautions were taken, with frequencies of perhaps 300,000 cycles. I had the benefit of Mr. Aceves' assistance in obtaining the data given hereafter.

The bridge was arranged as shown in Figure 9. The source of power was a calibrated oscillator which was adjustable from 1,000 to 300,000 cycles a second. The output of this set was supplied to a potentiometer *P* from which a known, adjustable voltage could be supplied to the bridge. Knowing the impedance of the bridge arms, the voltage impressed on the device to be tested can easily be calculated. The detector set was connected as shown and consisted of an oscillating tube circuit, the output of this circuit being supplied to a three-stage low-frequency amplifier, equipped with a loud speaking horn so that the bridge operator could make the balance without having the inconvenience of wearing a telephone receiver all the time. The low frequency amplifier had a voltage increase of about 1,000 and the heterodyne gave considerable amplification, the amount depending on the conditions.

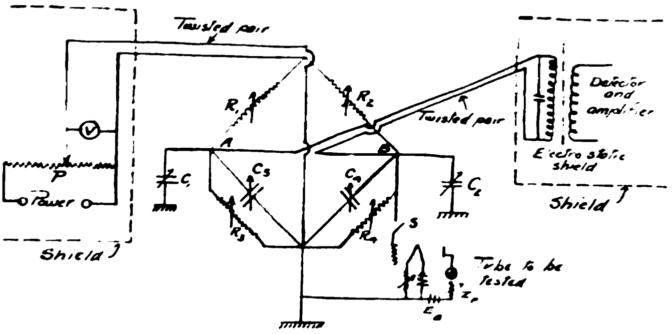

FIGURE 9

The power generating set and the detector set were placed far apart and each was completely enclosed in a suitably designed screening chamber. The power and detector lines leading to the bridge were twisted pairs covered with a grounded shield. The pair of wires leading to the detector did not connect to the oscillating tube circuit but connected to a coil which acted on the detector proper by magnetic induction. The electrostatic induction into the detector circuit was completely eliminated by a suitable shield which permitted magnetic induction but short-circuited electrostatic effects. This precaution is necessary because even when the bridge is balanced the line leading to the detector is excited by the power set as the middle points of the bridge, to which the detector is connected, are considerably above ground potential. The shielding was done so well that no signal was audible in the detector circuit unless the detector line was actually connected to the bridge.

The bridge itself was sufficiently well constructed that no errors due to high frequency effects could be found with currents of 300,000 cycles a second. It could probably be used at frequencies as high as 10^6 cycles a second without large error.

The two condensers C_1 and C_2 serve to correct for any distributed capacity in the bridge and its connections; with the (3) and (4) arms open, these two small condensers are adjusted so that the bridge is balanced, thus bringing the two points A and B to the same potential with respect to ground. If now the R_3, C_3, and R_4, C_4, arms are connected in and the bridge balanced by suitable adjustment of these four quantities, then the impedances of these arms (and of course the respective components of the impedances) are in proportion to the values of resistance in the ratio arms.

The radio frequency bridge seems to be most reliable if resistances of the order of a few hundred ohms are used in the ratio arms. Resistance units of about 100 ohms seem to have minimum error due to the combination of skin effect and internal capacity.

The tube to be tested was connected as indicated in Figure 9, the switch S serving to connect it or disconnect it as desired. The bridge having been balanced (with S open) by adjustment of C_4 and R_4, the switch S was closed; the capacity and conductance thus introduced into the bridge destroy the balance, and it is restored by decreasing C_4 and suitably changing R_4 (decrease or increase as the case may require). The amount of change in C_4 gives the capacity of the tube, and the required change in R_4 permits the calculation of the conductance of the tube.

The various types of tubes at present available were thus measured, with no filament current and no plate voltage, to determine what I call the geometrical capacity of the tube, that is, the capacity of the various parts with respect to one another without the presence of electrons or any of the various interactions which occur when the tube is connected in an operating circuit. By proper differential measurements the capacity of the leads, tube holders, etc., were eliminated so that the constants of the tubes themselves were obtained. When in operation on a circuit, the capacities will be somewhat greater due to the bases, connecting wires and so on.

The tubes tested had the following approximate ratings:

Number	Filament current	Plate volts	Plate current	Type of Filament	Intended service
1	1.1	20	6×10^{-4}	Oxide	Detector and Amplifier
2	1.1	20	4×10^{-4}	Tungsten	Detector and Amplifier
3	1.3	130	7×10^{-4}	Oxide	Amplifier
4	1.75	350	5×10^{-2}	Tungsten	Power
5	1.35	300	4×10^{-2}	Oxide	Power
6	6.5	500	15×10^{-2}	Tungsten	Power
7	3.65	1,000	25×10^{-2}	Tungsten	Power

Several tubes of each kind were tested giving typical results as tabulated herewith; the capacities being in 10^{-12} farads.

	1	2	3	4	5	6	7
Grid to filament, plate free	10.4	6.4	6.8	5.6	7.6	8.0	55.6
Grid to plate, filament free	14.4	4.4	7.6	3.0	8.4	8.0	22.0
Grid to plate and filament, these being connected together	17.0	7.2	12.4	7.2	11.2	10.2	69.6

These values are good to about 0.5 of a micro-microfarad, it not being thought while to work for a greater precision. The conductance was in all cases negligible, being only a small fraction of a micromho.

These magnitudes of capacity are such that they might well be neglected except for very high frequency circuits or for very carefully designed high frequency amplifiers. It seemed quite possible that when the space between the grid was saturated with electrons, these values of capacity might be greatly increased so the next set of values were taken to see whether the tube had more capacity when the filament was heated than when the tube was cold. It seems plausible to believe that these electrons, between the two plates of the condenser increase the capacity in the same way that it would be increased if a piece of metal (having electrons free to move) were placed in the space between the grid and filament. For all the tubes tested the increase in capacity due to heating the filament was less than one micro-microfarad. There was distinct evidence that the effect was present but as the purpose of the research was to discover in tubes, capacities of sufficient magnitude to effect their behaviour in ordinary circuits I did not try to refine the measurements really to investigate this point.

As soon as the filament was heated the conductance increased very rapidly, depending upon the value of the filament current, amount of negative voltage used on the grid, magnitude of the plate voltage, and the magnitude of the voltage impressed on the grid for the purpose of balancing the bridge. Figures 10, 11, 12, and 13 show the effect of these various quantities for one of the tubes used; the others showed similar effects.

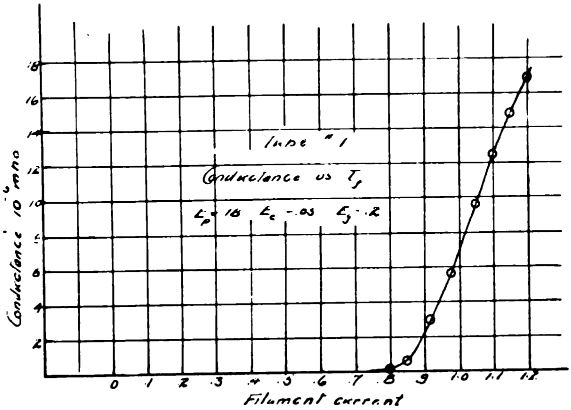

FIGURE 10

The next condition investigated was the plate circuit impedance, this promising very considerable effect as an elementary analysis showed that the capacity between the grid and ground might be expected to vary a great deal as the impedance in the plate circuit was varied. Figure 14 serves to show the elements of the problem; the filament is ordinarily grounded and, as the plate is also grounded by its connection thru the plate im-

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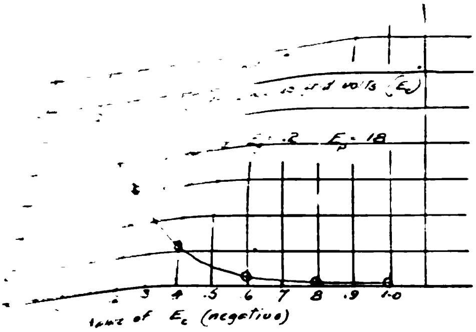


FIGURE 11

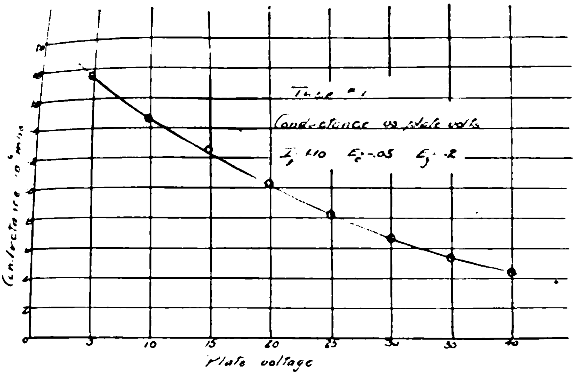


FIGURE 12

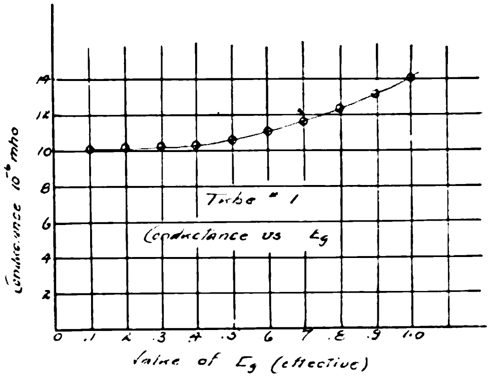


FIGURE 13

pedance Z_p and the battery E_B, any alternating emf. impressed between the grid and filament must produce a charging current sufficient not only to charge the condenser consisting of the grid and filament, but also to supply that for the condenser made up of the grid and plate. Suppose a voltage E_g is impressed between

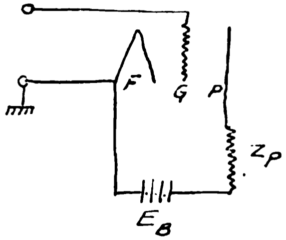


FIGURE 14

the grid and filament; there must be a charging current equal to $I=2\pi f C_{g-f} E_g$ to supply the electrostatic energy for this condenser. Now due to the amplifying action of the tube, the voltage E_g from grid to filament produces a voltage between grid and plate nearly equal to $(\mu+1)$ E_g, μ being the voltage amplification factor of the tube and its attached circuit. The charging current for the condenser, grid-plate, must be supplied from the source supplying E_g, so that the total charging current required by the input circuit is given by the equation

$$I=2\pi f E_g\,[C_{g-f}+(\mu+1)\,C_{g-p}]$$

which means that the effective capacity between grid and ground is

$$C=C_{g-f}+(\mu+1)\,C_{g-p}.$$

With this idea in mind it is seen at once that the capacity of the input circuit is much greater than might be supposed, and that it varies moreover with any factor which changes the amplification factor.

It might seem that the factor μ instead of $(\mu+1)$ should be used when taking into account the grid-plate capacity but it must be remembered that μ is the ratio of amplitudes of the plate-filament and grid-filament voltages (alternating current values, of course) and that the phase of these two voltages differs by nearly 180 degrees; this makes the potential difference of the grid-plate nearly equal to $(\mu+1)\ E_g$. If the two voltages were exactly 180 degrees apart the expression $(\mu+1)$ would be correct.

Moreover, because of the reaction of the plate-filament field on the potential of the grid, it might be expected that the conductance curve of grid to ground might also show very different characteristics than it does without the reaction of the plate field. A complete series of results were taken therefore to investigate these two effects and the expectations were fully justified.

Figure 15 shows the capacity and conductance of tube number 1 as the resistance of the plate circuit was varied; on the same curve sheet is shown the value of the voltage amplification factor of the tube for the various plate circuit resistances.

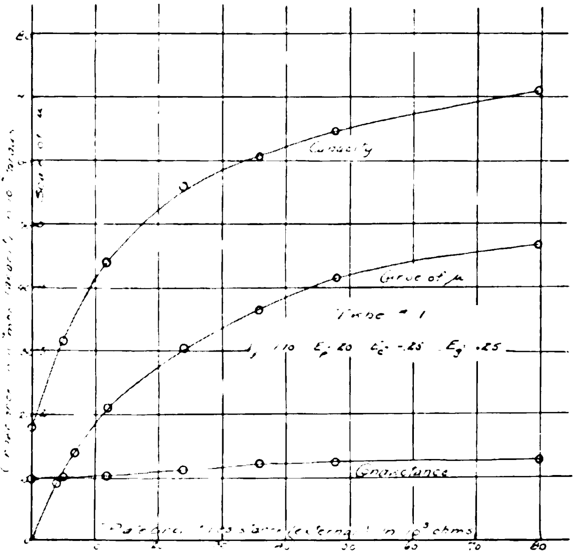


FIGURE 15

It is seen that the capacity of the grid to ground increases from 17 $\mu\mu$ farads to 71 $\mu\mu$ farads as the plate circuit resistance was increased from zero to 80 kilohms. As the capacity of the grid to filament is 10.4 and that of the grid to plate is 14.4, it might be expected that for the plate circuit resistance of 80 kilohms, the capacity should be equal to $[10.4+(4.65+1)14.4]=91.6\ \mu\mu$ farads. This calculation would undoubtedly give a very close approximation were it not for the fact that a considerable part of the plate-grid capacity is mutual with that of the grid-filament capacity, this mutual capacity having the effect of decreasing the grid-plate capacity.

The conductance of the grid-ground circuit gradually increased as the plate circuit resistance was increased.

In Figure 16 are shown the capacity and conductance of the grid-ground circuit of tube number 1 when reactance was used in the plate circuit instead of resistance. In this case the predicted increase in capacity is more nearly realized than when resistance is used in the plate circuit. Thus, with a reactance in the plate circuit of 50 kilohms (the value of μ being 4.2) the calculated capacity is 85.2 whereas the experimental value is 81.5 $\mu\mu$ farads.

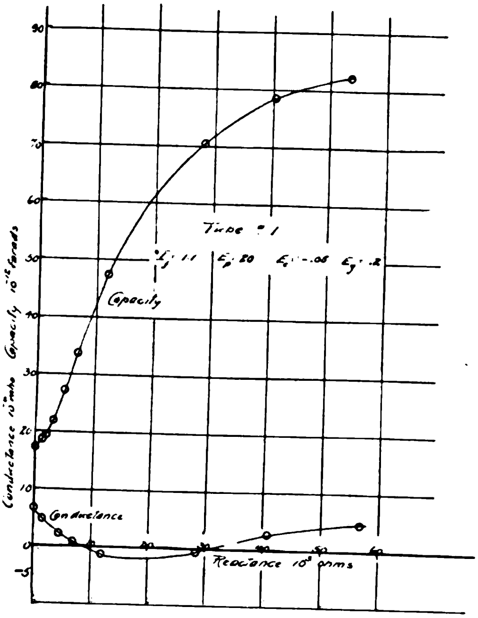


FIGURE 16

That any capacity present between plate and grid and which is not in the field of the grid-filament capacity is increased by the factor $(\mu+1)$ was proved by actually connecting a capacity of 20 $\mu\mu$ farads across the plate-grid terminals of the tube and noting the increase in the apparent capacity of the grid-ground circuit. It increased by 102 $\mu\mu$ farads.

The conductance of the input circuit of the tube becomes negative for certain values of plate circuit reactance, showing that, to a certain extent, the reaction of the plate circuit on the grid circuit is such as to supply energy from the plate circuit to the grid circuit, with no other coupling than that in the tube itself. It is of course known that under certain conditions the capacity coupling in the tube itself is sufficient to maintain oscillations, a condition which can be readily predicted from the conductance curve given.

Figures 17-21 show the characteristic curves of some of the other types of tubes tested. It is seen that the same general shape holds for all three electrode tubes, the difference being one of degree only. The capacity of the grid-ground circuit of the tube, when the tube is operating with a normal amount of resistance or reactance in the plate circuit, is from 5 to 10 times as much as the geometrical capacity of this circuit and the amount of this increase is controlled by the capacity between the grid and plate.

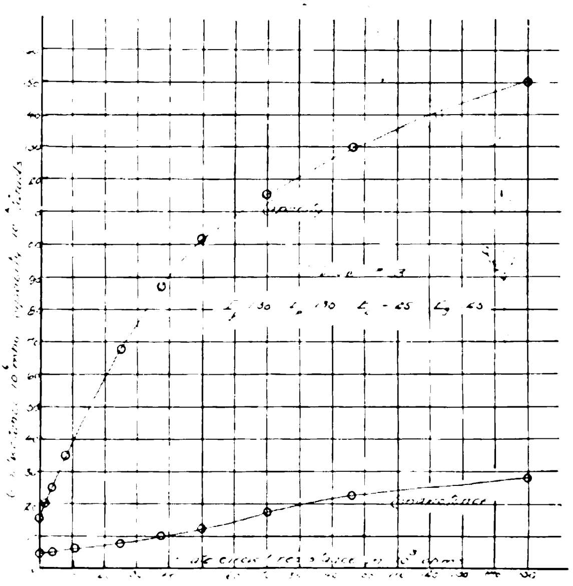


FIGURE 17

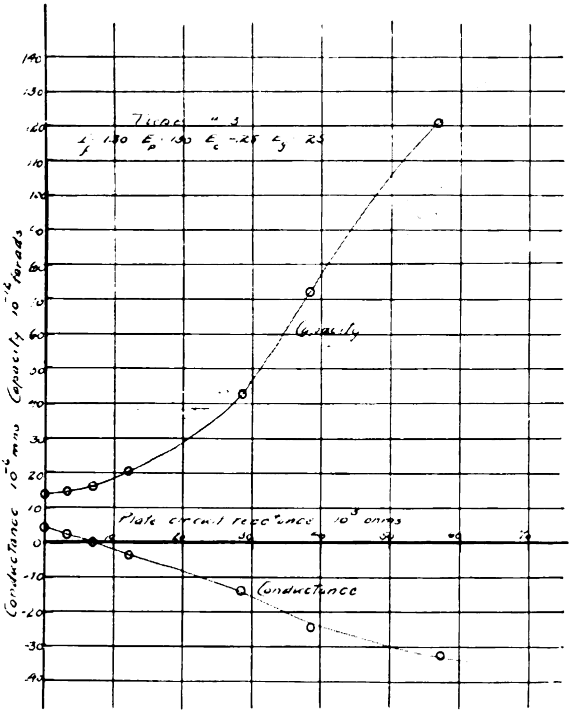


FIGURE 18

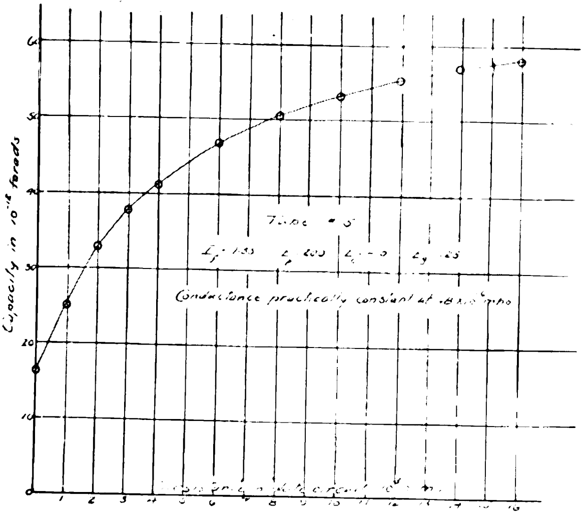


FIGURE 19

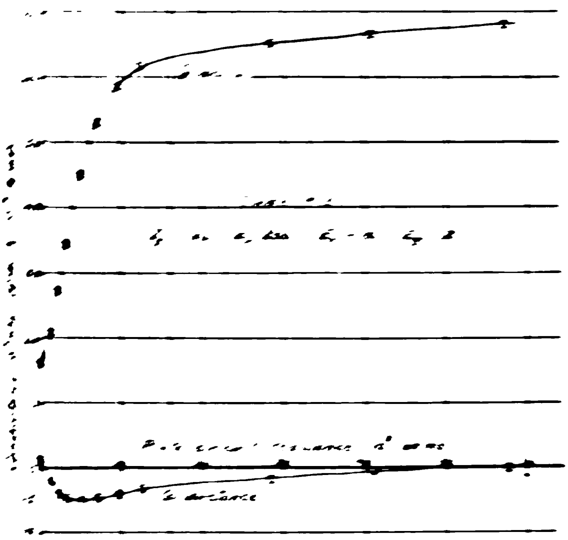


FIGURE 20

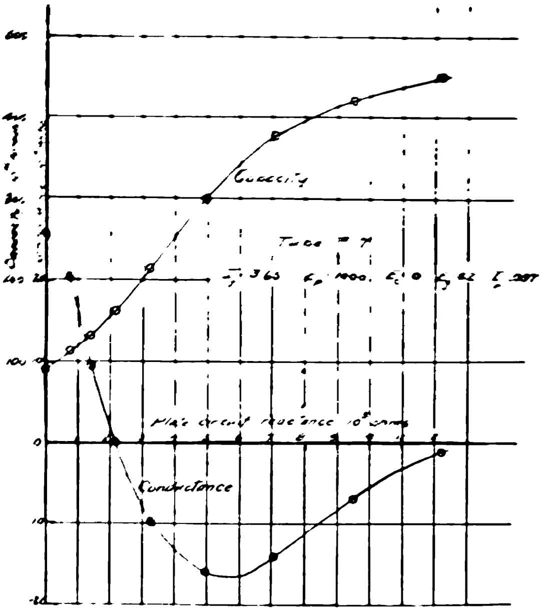
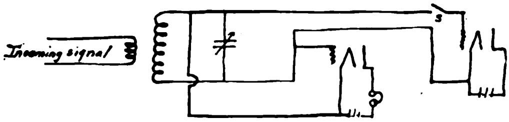


FIGURE 21

Detecting Action of a Tube Without Grid Condenser

The suggestion has been made* that when a three electrode tube is used as a detector of radio frequency oscillations without the use of a condenser in series with the grid, the detecting action is due to the asymmetrical character of the conductance of the grid-ground circuit, this asymmetry in conductance producing an asymmetry in the voltage wave in the closed receiving circuit. Calculation shows that this is at least a plausible explanation of the action but so many assumptions must be made in calculating the problem that experimental verification seemed wise. A circuit was arranged as shown in Figure 22; two detecting circuits being arranged to receive the same signal. The second tube circuit could be connected or not as desired by the switch S. It will be seen that the connection of the grid-filament circuit is so arranged that any asymmetry caused by the peculiar conductance value of the grid-filament of one tube is just neutralized by the action of the other tube. The audibility of the signal was measured with and without the second tube connected and no appreciable difference in the audibility under the two conditions could be detected, altho the tuning was not quite as sharp with two tubes connected as with only one. This effect is of course to be expected. The test seems to prove that the asymmetry of the grid-filament circuit of the tube plays an unimportant role in the detecting action of the tube.



Each tube with its A and B batteries was completely insulated from ground

FIGURE 22

Marcellus Hartley Research Laboratory,
Columbia University, June 4, 1919.

SUMMARY: The value of the exponent connecting plate current and plate and grid potentials is experimentally investigated for a number of tubes. It is found to be markedly different for different tubes, and for the same tube at different plate voltages. Particularly is this the case for low plate voltage tubes.

---

* Bown, "Physical Review," volume 10, number 3, page 257.

After discussing briefly distortion in amplifying tubes, the capacity and conductance within tubes is considered. Experimentally determined data governing these quantities are given, these being obtained on a carefully designed radio frequency Wheatstone bridge.

An experiment on the detecting action of a tube without grid condenser seems to prove that grid-to-filament circuit asymmetry of characteristic is not responsible for the detecting action.

PROCEEDINGS OF

The Institute of Radio Engineers

Volume 8 AUGUST, 1920 Number 4

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CORRECTION: In THE PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS for April, 1920; Volume 8, Number 2, page 165, the formula on line 3 should read:

$$C=\left(4\sqrt{a}+0.885\frac{a}{h}\right)\left(1+0.015\frac{l}{b}\right)\cdot 10^{-6}$$

and in the Table 1, the seventh column of the second portion of the Table should be headed $\frac{l}{b}$

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PUBLISHED BY
THE INSTITUTE OF RADIO ENGINEERS, INC.
THE COLLEGE OF THE CITY OF NEW YORK

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

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TRANS-OCEANIC RADIO COMMUNICATION*

By

ERNST F. W. ALEXANDERSON

(CHIEF ENGINEER, RADIO CORPORATION OF AMERICA, NEW YORK)

It has already become generally known that a new highway for world traffic has been opened up thru the development of trans-atlantic radio communication. It is now a matter of history that radio was largely used for communication between the United States and armies in Europe, and that the Great War was brought to a close by negotiations conducted by radio which led to the Armistice. Now, we are ready for an international commerce of unprecedented scope, but lack adequate means for communication.

The recent achievements of radio technique have become common knowledge, and the world has now turned to this new method of communication clamoring that it step in and save the day. This is a condition which places a serious responsibility upon radio engineers. Fortunately, the technique has emerged from the cloud of mystery that used to surround it, and we are in position to treat the problem coolly and scientifically like any other problem in electrical engineering. However, it must not be inferred that the task is an easy one, if the radio technique is to fulfill all the hopes which are placed on it.

It has been demonstrated during the war period that trans-oceanic communication has become thoroly reliable, every day in the year, and practically every hour in the day. Thus far, we can say, that the problem is solved. But a second question will be raised: What volume of traffic can be carried by the means at our disposal at the present time, and what is the relation of this radio traffic to the world traffic of to-day and to the world traffic of the future? The facts of the case are briefly the following:

Experience has shown that the wave lengths which are most suited for trans-oceanic communication lie between 12,000 and 17,000 meters. This "space in the ether" has already been taken

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* Received by the Editor, September 8, 1919. Presented before a joint meeting of THE AMERICAN INSTITUTE OF ELECTRICAL ENGINEERS and THE INSTITUTE OF RADIO ENGINEERS, New York, October 1, 1919.

up by five first-class transmitting stations which, during the war period and up to the present time, have been in continuous service for trans-atlantic communication. Of these stations, two are in the United States, one in England, one in France and one in Germany. By extending the range of wave lengths down to 10,000 and up to 20,000 meters, and following the same system of intervals there would be room for about seven more stations or a total of twelve first class transmitting stations.

A first class station has such radiating power that its messages can be received in all parts of the world. This is one of the advantages of radio communication; but it implies that if such a station is to be used to full advantage it must have a "right of way" for its wave length over the whole world. Thus if we look at the matter pessimistically, without allowance for the improvements that further engineering developments are likely to bring, it would look as if the capacity of the world for first class radio stations would be about twelve. The rate of transmission at the present time from these stations is about twenty words a minute and it would thus be easy to figure out that the capacity of radio for handling any considerable portion of world communication would be totally inadequate.

The other side of the picture, which the radio engineer of the future must study carefully and closely, should indicate the technical possibilities for improving the situation. The tendency of present day developments points to the following means for expansion of radio traffic:

1. Increase in speed of transmission.
2. Improved selectivity based on the direction of the wave.
3. Improved selectivity making possible closer spacing of wave lengths.

As a basis of discussing the general situation it may be here stated as simple facts:

1. That signals have been transmitted and received at considerably more than 100 words a minute.
2. That signals have been received from Europe while an American high power station within comparatively short distance was radiating on the same wave length.
3. That it has proven practical to separate radio signals differing in wave length considerably less than 1 per cent.

Based on these facts, it is probable that the transmitting speed in the future will average 100 words a minute instead of 20 words a minute.

That the selectivity for direction of the waves will multiply by five the number of stations that may be operated on one wave length; and that the selectivity with reference to wave length will be improved so that the wave lengths of messages will be within 1 per cent. of each other, instead of 7 per cent., which is the spacing of the stations at present.

These prospects, taken in combination, give us an optimistic picture in which the possible capacity for trans-oceanic radio traffic of the world is 175 times as great as it is with the practice of to-day.

To claim that the traffic capacity could immediately be increased, nearly 200 times, would be an exaggeration, because the different improvements which have been made may partly conflict with the execution of each other if they are to be used simultaneously. This optimistic picture is, therefore, to be regarded as a goal—perhaps never to be reached, but it points the way for almost unlimited possibilities for progress by continued engineering efforts.

In order to avail ourselves of these improvements simultaneously, the transmitted wave must be a continuous wave which does not "spill over" with harmonics, decrements or variations, into the range of wave length assigned for other communication. When wave lengths are spaced 1 per cent. apart, each wave must be sufficiently pure so as to have no objectionable components outside the limits of one-half of one per cent. While rules to this effect should be rigidly enforced, it must be appreciated that there are certain fundamental limitations.

HIGH SPEED SIGNALLING

Modern trans-oceanic radio signalling is conducted by means of continuous waves. It must be appreciated, however, that signalling by an absolutely continuous wave would be impossible, because the making of dots and dashes introduces increments and decrements in the radiation. It will be shown that a repetition of increments and decrements can be resolved into a group of closely adjacent continuous waves. This agrees with the well-known fact that the tuning of a wave with a decrement is known as a broad tuning. To illustrate this point we may take as a basis a signal at 100 words a minute (5 letters per word). If it is assumed that the increments and decrements in making the dots are not sharp interruptions but a continued variation by sine wave curves as indicated in Figure 1, it is found by analysis that such a wave may be resolved into a group of continuous wave components within the limits of 40 cycles above and below the average. This is the theoretical minimum width of the band of wave lengths which are necessary to transmit 100 words per minute. If the dots are defined by sharper interruptions the wave becomes still "broader" and it would not be unreasonable to say that the minimum practical band of wave lengths for 100 words per minute is 100 cycles on each side of the fundamental. This would make possible a spacing of the waves 1 per cent. apart when using a wave length of approximately 15,000 meters. Messages at a higher rate of speed will occupy a correspondingly wider "space in ether."

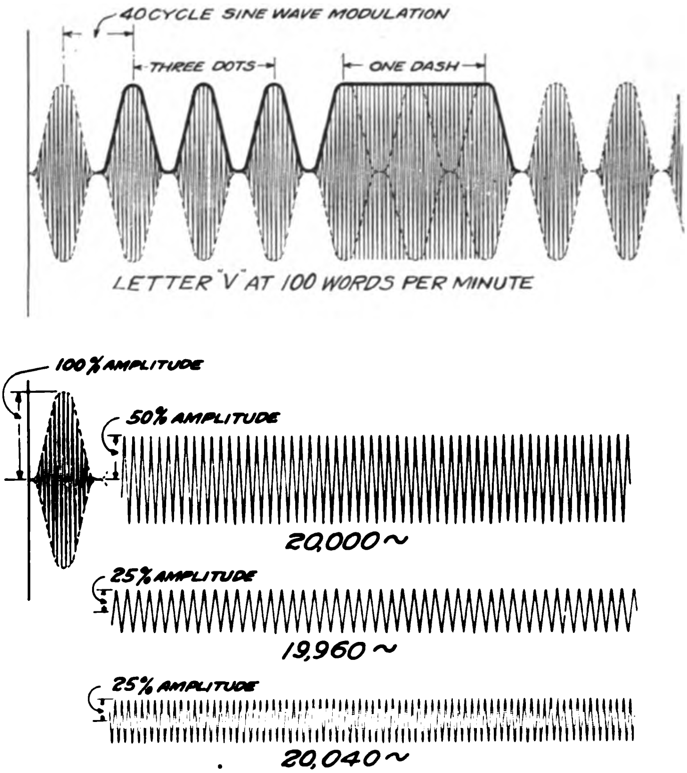


FIGURE 1—Method of Resolving Sine Wave Modulation Into Three Continuous Waves

As a conclusion from this analysis it may thus be said that an increase of the speed to 100 words a minute and increase of messages to a spacing 1 per cent. apart may be accomplished simultaneously, provided that waves are used of such a character and modulation that they contain no radiation except the one needed to accomplish the intended purpose.

Directive Reception

The second means for increasing radio traffic consists in improving the selectivity by taking advantage of the direction of the waves. The author's paper printed in the Proceedings of the Institute of Radio Engineers, August, 1919, described a receiver referred to as the "barrage receiver" which was used to demonstrate directive reception. There are several other types of receiving devices, developed by the United States Navy and other investigators, which accomplish substantially the same purpose. The broad principle underlying all these directional receiving devices is the one discovered by Bellini and Tosi, and generally referred to under their names.

In the United States Navy's tests of the "barrage receiver," it was proven that it is possible to carry on simultaneous two-way communication on exactly the same wave length. If this method of directive reception is carried out consistently in a world system of communication, it may be assumed that transmitting stations operating on the same wave length may be located approximately as shown in the map in Figure 2—one in Europe, one on the American east coast, one on the American west coast, one in the Far East, and one in South America. The American receiving station for European signals in such a system should be located east of the American Atlantic trans-

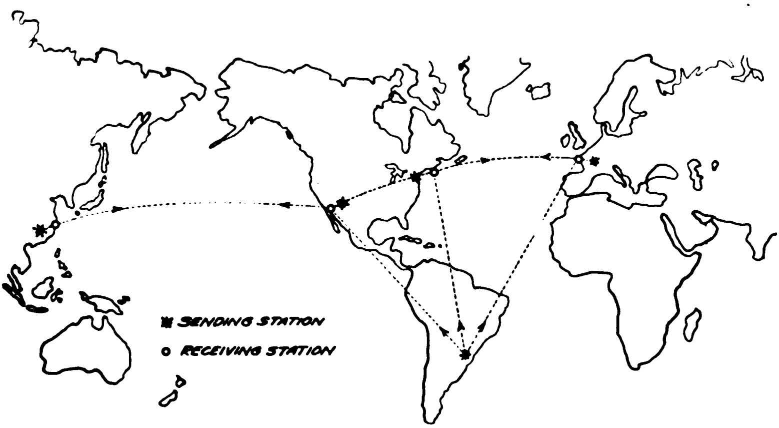


Figure 2—Proposed Simultaneous Radio Transmissions on the Same Wave Length

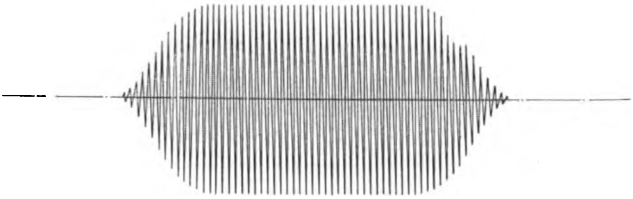
mitting station, and in line with the Pacific transmitting station. Thus messages from the two American transmitting stations could both be simultaneously neutralized in the American receiving station by a "barrage receiver," while signals on the same wave length are received from Europe. Interference from the South American station may be neutralized by the use of a double barrage system, while the station in the Far East, tho it may not be exactly in line with the two others, would be sufficiently near the general direction so that, considering the great distance, it would not cause interference.

If this communication system is to be duplicated on a number of other wave lengths, the practical conclusion follows that the transmitting stations as well as receiving stations for each district should be grouped in centers, and these centers located relatively so as to make the directive neutralization as effective as possible. The neutralization of several transmitting stations simultaneously may not always work out as first designed. It has been shown by investigations of the Navy Department that the radio waves do not always follow straight lines and not always the same path. However, discrepancies from such origin may again be overcome by further extending the principle of neutralizing waves from several directions simultaneously.

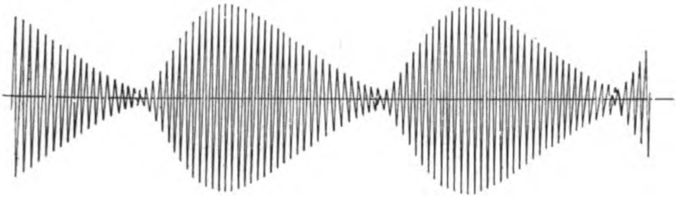
CLOSER SPACING OF WAVE LENGTHS

The method of increasing traffic capacity by closer spacing of wave lengths has great possibilities. It has been shown that the selectivity with reference to wave length can be greatly increased by several successive tunings in either the radio, the audio or some intermediate circuit. It is thus entirely practical to receive either by ear or by photographic records, signals which are considerably less than 100 cycles apart. The theoretical limits for such selectivity in connection with high speed transmission are defined below. As illustrated in Figure 1, a high speed message is not a single continuous wave, but a band of wave lengths—100 words per minute occupying a space of about 200 cycles. Therefore the same degree of selectivity is not to be expected with a high speed signal if the interfering radiation is of considerable intensity.

Speaking in terms more familiar to practical radio operators it may be said that high speed telegraph signals assume to some degree the objectionable characteristics of a spark signal. As an illustration of this, Figure 3 shows the relative decrements in a continuous high speed telegraph wave and a spark wave,



WAVE TRAIN SHOWING ACTUAL NUMBER OF RADIO CYCLES IN A DOT OF THE TELEGRAPH CODE TRANSMITTED AT THE RATE OF 500 WORDS PER MINUTE AT 15,000 METERS WAVE LENGTH



A Spark Wave at 400 Sparks Per Second

FIGURE 3—Decrement of High Speed Signal Compared with Spark Wave

showing that the increments and decrements of the continuous wave signal at 500 words per minute are about equal to a spark wave of 400 sparks per second. These illustrations apply to wave lengths of about 15,000 meters.

THE RADIO TRANSMITTING SYSTEM

Several types of radio transmitting systems are at present in use with a high degree of success. The descriptive matter in this paper will, however, be confined to the system for which the author is responsible, as represented by the Naval Radio Station at New Brunswick, New Jersey.

Generally speaking, any radio transmitting system consists of three essential elements:

1. The generator of radio frequency energy.
2. The modulating system, whereby the energy is controlled so as to produce the dots and dashes of the telegraph code or the modulations of the human voice.
3. The antenna or radiating system.

GENERATING SYSTEM

There are four types of generating systems of radio frequency energy in use at the present time.

1. The spark or impulse generator.
2. The Poulsen arc generator.
3. The radio frequency alternator.
4. The vacuum tube oscillator.

The system which will be described is of the type employing a radio frequency alternator. The installation in New Brunswick contains a 50-kilowatt alternator shown in Figure 4 which was operated for some time for experimental purposes with radio telephony at a wave length of 8,000 meters, and later in transatlantic telegraph service at 9,300 meters.

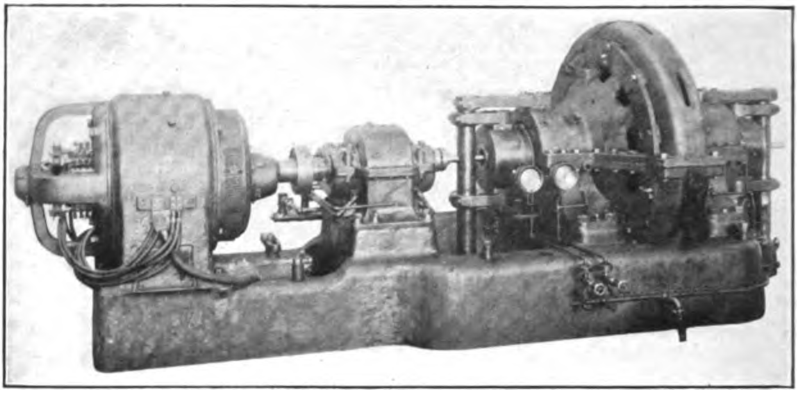


FIGURE 4—50-Kw. 50,000 Cycle Alternator

A larger equipment which has been in continuous service for the last year consists of a 200-kilowatt alternator shown on Figures 5, 6, and 7. Figure 5 shows the machine partly assem-

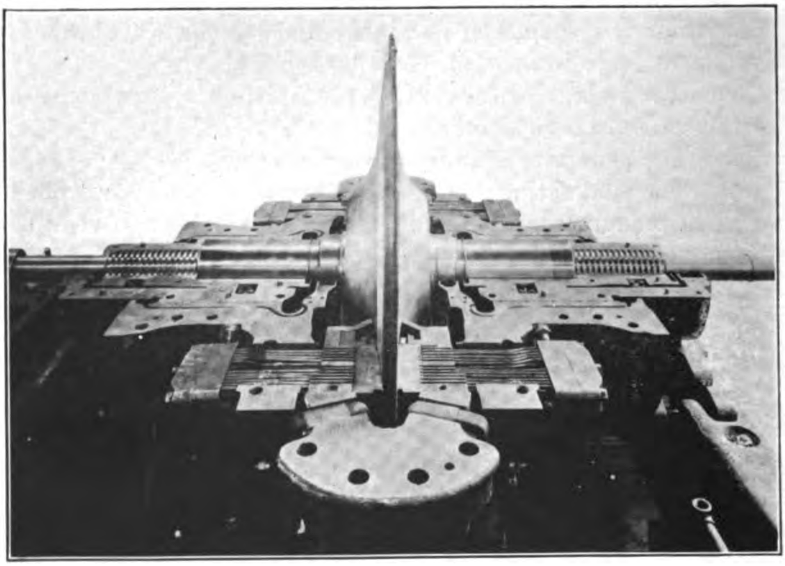


FIGURE 5—200-Kw. Radio Frequency Alternator

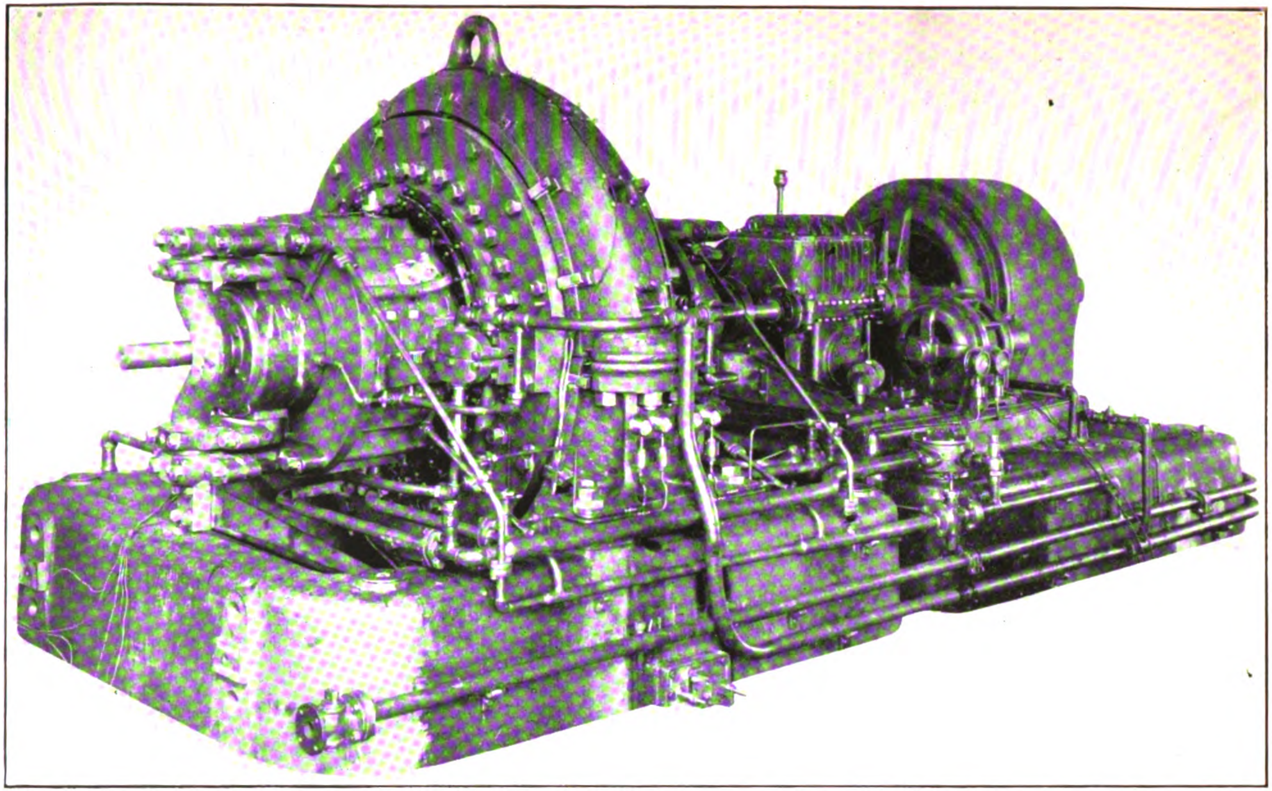


FIGURE 6—200-Kw. Radio Frequency Alternator

bled, the rotor consisting of a solid steel disc. The spaces between the polar projections are filled with non-magnetic material so as to present a smooth surface and thereby reduce air friction to a minimum. The disc runs between the two laminated armatures, which are cooled by water pipes, as shown in the photograph. The armature winding which consists of wire wound back and forth in straight open slots, is divided in 64 sections, each section generating about 100 volts and 30 amperes. The current generated by these 64 windings is collected in the air-core transformer mounted on the top of the machine (Figure 7). This transformer has 64 independent primary windings corresponding to the armature windings. The single secondary winding of the transformer delivers the complete output of the alternator. This collecting transformer is thus to be considered as an integral part of the generating unit; and for all purposes of calculation, the characteristics of the generating unit, such as electromotive force and current, are given as delivered from this secondary winding. At full output the alternator delivers 100 amperes at an electromotive force of 2.000 volts. It can thus be seen that the alternator is designed for a load resistance of 20 ohms. However, the same machine might be adapted for any other load resistance by selecting a different number of turns in the secondary of the collecting transformer. The reason why this particular machine is designed for a high voltage and low current will be given later in discussion of the new type of antenna with which it is used.

The 200-kilowatt alternator when operated at the New Brunswick wave length of 13.600 meters runs at a speed of 2,170 r.p.m. It is driven by an induction motor thru a gear of a ratio of 2.97:1. When the radio frequency alternator is used as a source of radiation, the wave length is determined directly by the rotative speed of the machine. Thus, obviously, it is important that the rotative speed should be as nearly absolutely constant as it is possible to make it. An important accessory of the alternator set is therefore the speed regulator. The 50-kilowatt alternator set shown on Figure 3 is driven by a direct current motor whereas, the 200-kilowatt set is driven by an induction motor of the slip ring type. The 50-kilowatt set was equipped by a direct current motor because the problem of speed regulation of that type of motor is somewhat easier. Induction motors were, however, decided upon for the later types because alternating current power is more easily available in most localities.

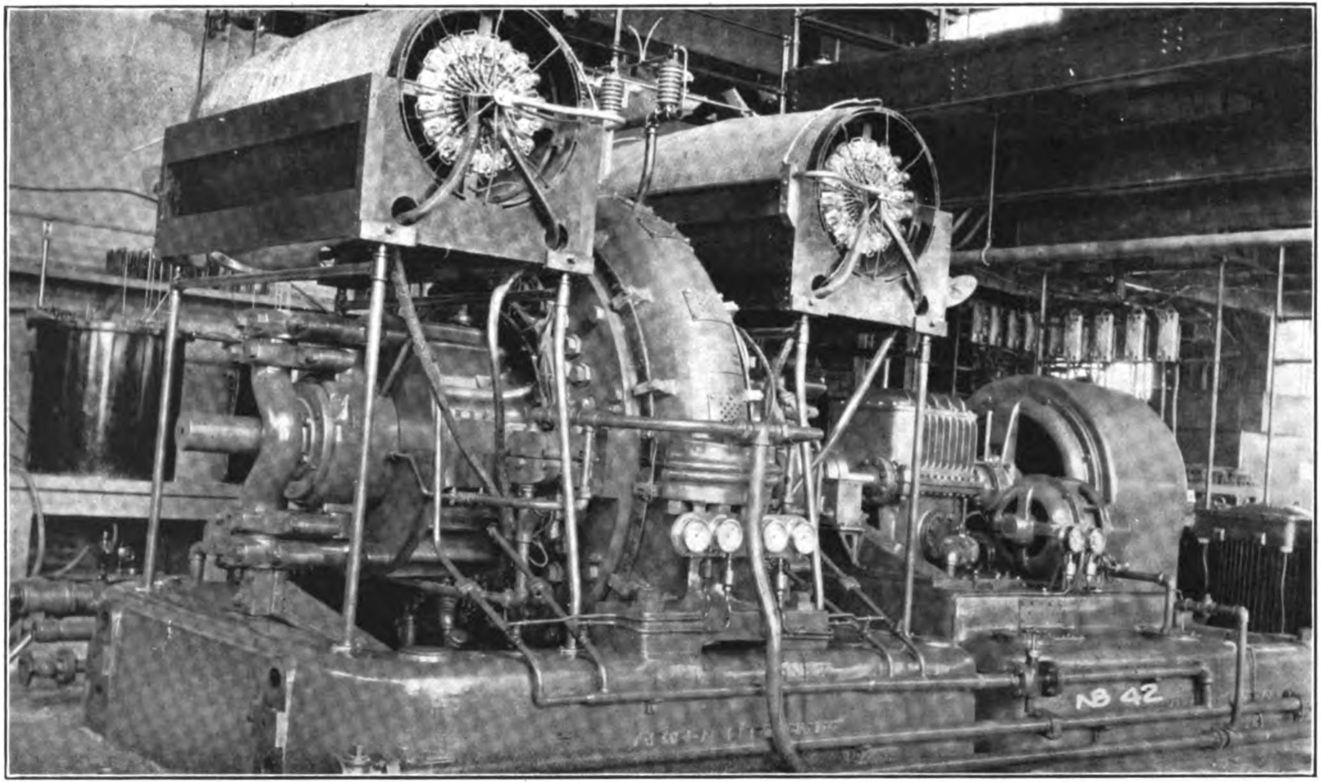


FIGURE 7—200-Kw. Alternator Set as Installed in the Naval Radio Station, New Brunswick, N. J.

SPEED REGULATOR

The speed regulator consists of a speed-determining element and a power-controlling element. The speed-determining element is a resonant radio frequency circuit fed by one of the 64 alternator windings which is set aside for that purpose. The oscillating energy of this radio frequency circuit is associated by magnetic couplings with a rectifying circuit in which the radio frequency energy is changed into direct current. This rectified current in turn, actuates the controlling magnet of a vibrating regulator of the type that is generally used for voltage regulation in power stations. When the driving motor is a direct current motor, it is easy to see how this vibrating regulator may be made to control the speed by regulating the voltage of the power supply to the motor. In order to accomplish the same object with an induction motor some new features have been introduced.

An ordinary induction motor is operated at constant potential. When the motor runs light, it draws from the line a magnetizing current which is almost wattless. Thus it operates at a low power factor. When the motor is fully loaded, it draws power at a high power factor—a motor of the type used having a power factor of 90 per cent.

When the New Brunswick station was adjusted for operation, it was found that a wave length was desired which required the induction motor to work at 19 per cent. slip. The rheostat in the secondary of the motor could easily be adjusted so that the motor would deliver the desired power with full load at 19 per cent. slip. However, inasmuch, as the output of the alternator varies continually with the making of dots and dashes of the telegraph code, the motor is alternately loaded and not loaded. The tendency would, therefore, be for the motor to speed up during the intervals. If the potential of the power supply to an induction motor is varied, the motor torque varies by the square of the voltage. It is furthermore easy to show, by the theory of the induction motor, that if a motor consumes power at 90 per cent. power factor at full load and the load is reduced to ¼ by the reduction of voltage to ½, the power factor will remain 90 per cent. In fact, it will always consume power at 90 per cent. power factor regardless of its load if the voltage supply is adjusted accordingly, so long as the secondary resistance remains constant and the speed remains constant.

Thus it may be said that the standard method of operating an induction motor is at constant potential and variable power factor. The method of operating the driving motor of the radio set may on the other hand be characterized as variable potential and constant power factor.

The problem which thus presented itself was to find means for varying the applied voltage in accordance with action of the speed determining element, and this has been done in the following way:

Between the motor and the power supply is introduced a choke coil with iron core, the permeability of which can be varied by saturation. The change in permeability is produced by a direct current which is controlled by a vibrating regulator. When the motor carries full load the iron core is saturated so that the choking effect is practically zero. At fractional load, the choking effect is automatically adjusted by the regulator so that the motor delivers at all times the power required to hold constant speed. The motor itself operates at all times at its maximum efficiency and power factor, but the power factor of the current drawn from the line varies with the load. Thus when the motor operates at ¼ load the power factor of the line is 45 per cent., while the power factor of the motor is 90 per cent. The circuits of the regulator are shown in Figure 8 and the photograph of the vibrator regulator in Figure 9.

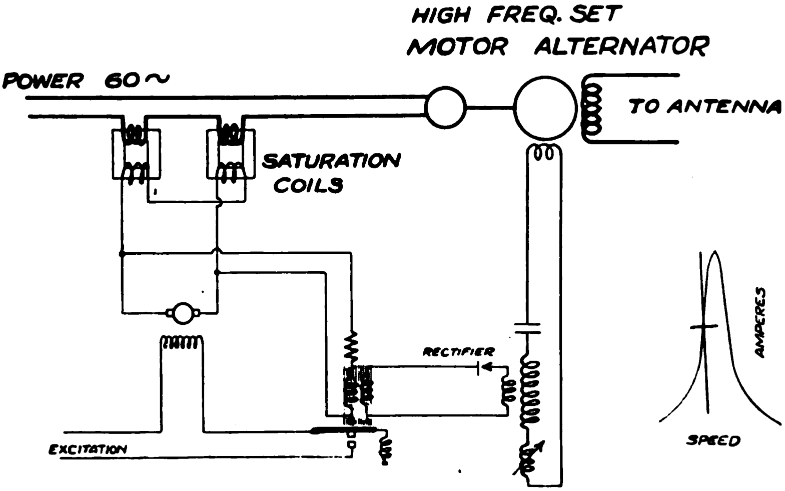


FIGURE 8—Speed Regulator Alternating Current Drive

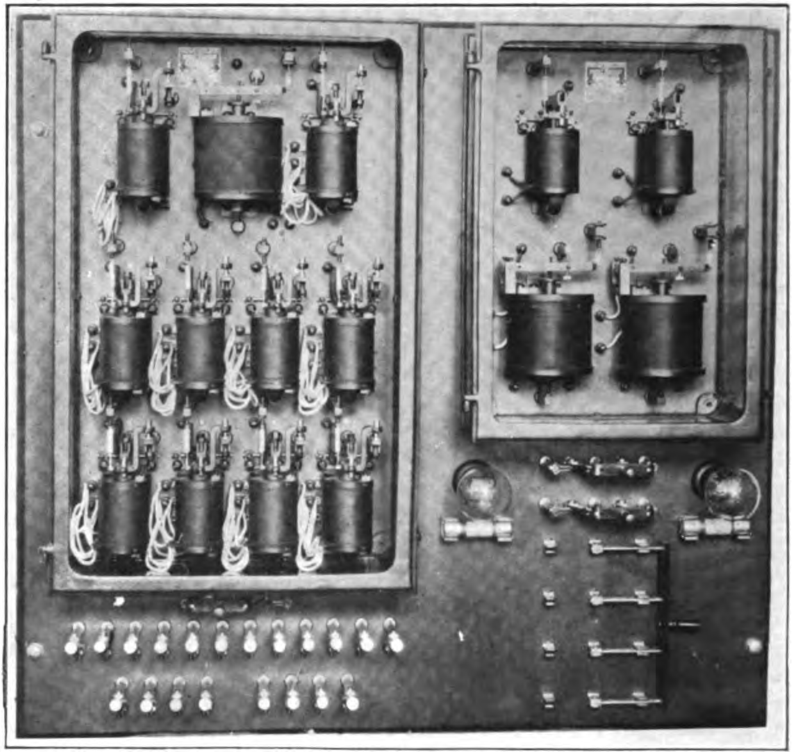


Figure 9—Combined Current and Voltage Regulators for Radio Frequency Alternator Control

Modulating System

The method of controlling radio frequency energy involves an apparatus which has become known as the "magnetic amplifier." This device is described in a paper by the author in the Proceedings of The Institute of Radio Engineers, January, 1916, and therefore needs to be referred to only briefly. The magnetic amplifier is a device which is physically of the nature of an oil-cooled transformer. The iron core which is made of fine laminations, is designed in such a way that the magnetic permeability of the iron core can be varied by magnetic saturation. By a special combination of tuned circuits as shown in Figure 10, it has become possible to separate the controlling current from the radio frequency current so that a comparatively weak current of a few amperes controls as many hundreds of amperes in the antenna. When the transmitting station is used for telegraphy, the magnetic amplifier is controlled by the telegraph relays which are a part of the wire telegraph system. During the war service, the telegraph key was operated in the centralized operating room of the Naval Communication Department in Washington. When the station is used for telephony the controlling current is an amplified telephone current.

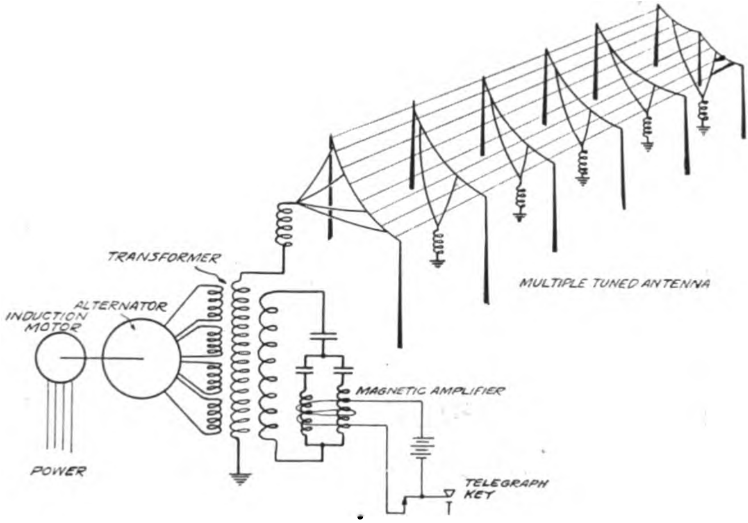


FIGURE 10

While the magnetic amplifier has proven to be a very satisfactory and reliable controlling device for ordinary telegraphy, its particular advantages are most prominent in high speed telegraphic transmission and telephonic transmission, on account of its instantaneous magnetic action without any arcing contacts. Figure 11 shows an oscillogram of radiation at 100 words per minute and a photographic record of reception at the same speed. Figure 12 shows the telephone modulation of the antenna current when Secretary Daniels was speaking over the telephone line from Washington, controlling the output from the New Brunswick station, thereby transmitting his voice to President Wilson's ship at sea.

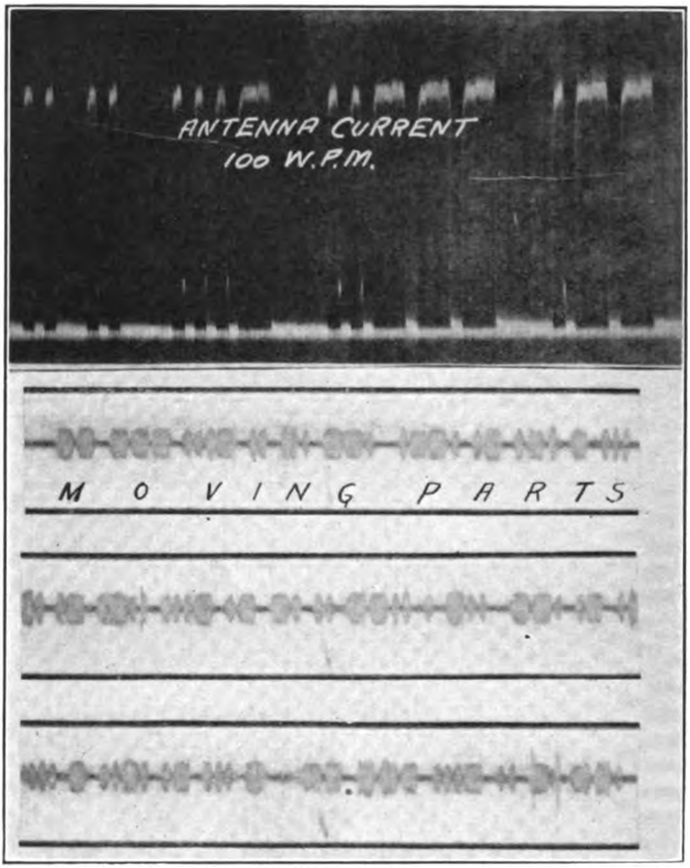


FIGURE 11—Oscillogram of Radiated Signal from New Brunswick; and Photograph of Received Signal from New Brunswick at 100 Words per Minute

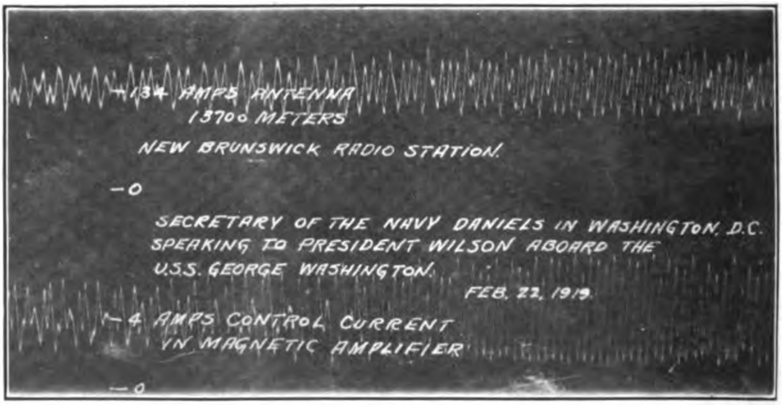


FIGURE 12

The Multiple Antenna

The antenna of the New Brunswick station represents a new departure in the method of radiation. The old antenna structure was originally one of the horizontal Marconi antenna, 5,000 feet (1,500 meters) long, 600 feet (180 meters) wide, supported on towers 400 feet (120 meters) high. The original antenna had a resistance of 3.8 ohms.

The antenna as operated now has a resistance of 0.5 ohm, distributed approximately as follows:

Radiation resistance	0.07	ohm
Tuning coils and insulation	0.10	"
Ground resistance	0.33	"
Total multiple resistance	0.5	ohm

The reduction in total resistance of the antenna is due to the reduction of the ground resistance. While the old antenna had one tuning coil located in one end, the new antenna has six tuning coils as shown on Figure 10 and Figure 13.

Theory of the Multiple Antenna

The theory of the multiple antenna can be explained in several ways. Without going into details, an explanation will be presented, giving the point of view which has proven most useful for general discussion.

For this purpose the multiple antenna may be considered as an aggregate of several antennas of the ordinary vertical type, each having its own tuning coil. When regarded in this way, the multiple antenna in New Brunswick is equivalent to six independent radiators placed 1,000 feet (305 meters) apart. The ground resistance of each of the radiators is 2 ohms, the coil resistance 0.6 ohm, and the radiation 0.07 ohm; making a total resistance of 2.67 ohms. Each of these radiators has an aerial 1,000 feet (305 meters) long and 600 feet (153 meters) wide, mounted at an average height of 300 feet (91.5 meters). A total resistance of about 2.67 ohms is, as a matter of fact, the resistance to be expected from an antenna of such dimensions and ordinary design. If one individual antenna, such as described, is operated with a radiation of 100 amperes, the energy consumption of the antenna would be 26.7 kilowatts, and the radiation efficiency would be 0.07 ohm divided by 2.67 ohms, which is 2.6 per cent. If it is desired to increase the radiation from 100 amperes to 600 amperes, the energy consumption would be 36 times as great, that is, 960 kilowatts. There is, however, another way to produce a radiation equivalent to 600 amperes. If six separate antennas of the dimensions described were built and each operated with a radiation of 100 amperes, each antenna would emit a system of waves proportional to 100 amperes radiation. If all the waves emitted by the six antennas were in phase, the amplitude of the combined wave would be six times as great as the wave emitted by one antenna. Thus the amplitude of the combined wave would be equal to the amplitude of the wave emitted by one antenna when operated

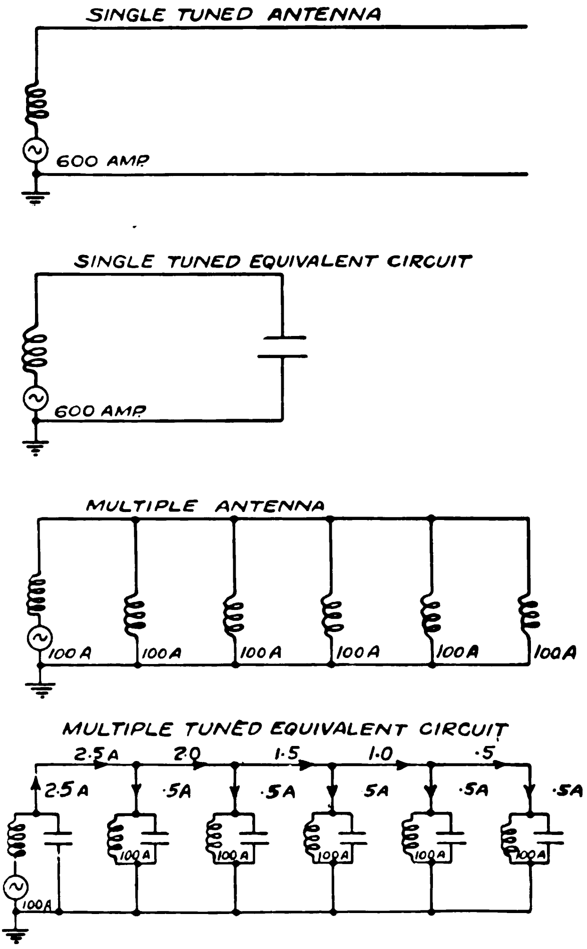


FIGURE 13

at 600 amperes. The energy consumption required for operating one antenna was 26.7 kilowatts, thus it might offhand be concluded that the energy consumption required for operating the six antennas simultaneously would be 160 kilowatts. This conclusion is not exactly correct because there is an interaction between the radiating effects of the different antennas resulting in a somewhat higher energy consumption. This might be expressed as follows:

The radiation resistance which is 0.07 ohm is common for all six antennas, whereas the ground and coil resistance of 2.6 ohms belongs to the different antennas individually. Thus the combined circuit of the multiple antenna can be represented by the common radiating resistance of 0.07 ohm connected in series with a group of six multiple resistances of 2.6 ohms each, so that the total current of 600 amperes flows thru the 0.07 ohm radiation resistances, while 100 amperes flow thru each of the 2.6 ohm resistances, which represent the ground and coil resistance of the individual antennas. Hence the total energy consumption of the combined antenna is found to be 180 kilowatts, out of which 155 kilowatts is ground coil and insulation loss and 25 kilowatts radiation. The radiation efficiency of the multiple antenna is thus 14 per cent. against the radiation efficiency of 2.6 per cent. for the individual antenna. The above calculation is made for the present operating wave of 13,600 meters. If, on the other hand, the same calculation is made for the wave length of 8,000 meters, which has been found more efficient for telephonic transmission, it is found that the radiation efficiency is 30 per cent., the energy being distributed as follows:

Radiation resistance	0.2	ohm
Coil and insulation	0.07	"
Ground resistance	0.33	"
Total multiple resistance	0.6	ohm

For the sake of convenience in dealing with the radiation from a multiple antenna, it has become practice to indicate the total radiation from the multiple antenna as the sum of the currents measured in the six ground connections. The equivalent multiple resistance of the antenna is then determined by the equation $I^2 R =$ energy consumption, where I is the sum of the currents in the ground connection.

It has been assumed above that each of the individual antennas is operated so that the waves sent out by the same are not only of the same frequency, but exactly of the same phase.

It remains to be shown how this is accomplished. Figure 13 shows the relation between the antenna and the multiple tuning coils and the alternator. As shown by the diagram, the alternator is connected in series with one of the six tuning coils. Arrows on the diagram indicate further how the oscillating currents and the energy currents are distributed. Six independent oscillating circuits are formed, the current in each ground connection corresponding to the charging current of the corresponding section of the aerial. If the antenna had been shock excited so that it continued to oscillate in the way indicated, no current would flow in the horizontal wires between the different sections of the antenna. However, in order to maintain continued oscillations a flow of energy must take place from the alternator, which is connected to one of the six tuning coils. When the antenna is operated with 100 amperes in each of the ground connections the energy consumption as shown above is 180 kilowatts, that is, each oscillating circuit consumes 30 kilowatts.

What actually takes place is the following: The tuning coil to which the alternator is connected transforms the energy of the alternator into a power supply at a potential of 60,000 volts, and each of the oscillating circuits draws energy from this power supply at that voltage. Thus the energy current consumed by each oscillating circuit is only 0.5 ampere. It can thus be seen that while the total oscillating current of the antenna is 600 amperes, the energy current which flows horizontally from the power source to the multiple oscillating circuits is only a total of 2.5 amperes. In other words the energy which is delivered by the first tuning coil in the form of 100 amperes at 1,800 volts is transformed by the first oscillating circuit and distributed as in a transmission line from which 0.5 ampere at 60,000 volts is drawn in five places. The analogy between the multiple antenna and a high tension power distribution system is thus apparent.

This point of view is a departure from the conventional theory of radiation; but it must be remembered that there was a time in the development of electric power technique when the introduction of the high tension multiple distribution system was a radical departure.

Directive Radiation

The multiple antenna as described in its simplest form is adjusted so that the radiation from each of the individual oscillators is in phase. If, however, the antenna dimensions are so chosen that the phase displacement of the travelling wave between the different radiators becomes an essential factor, it is possible to obtain directive radiation. The radiated wave will

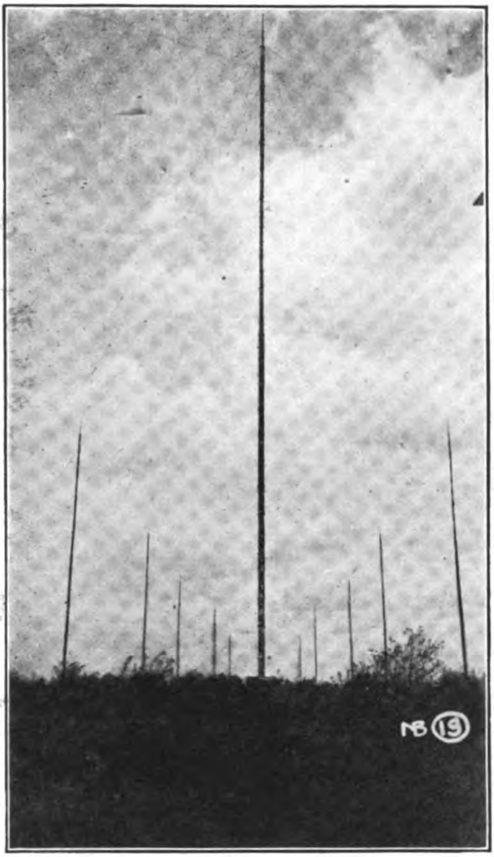


Figure 14—Perspective of Antenna

then not be a simple circular wave, but an interference pattern which may be treated like the corresponding phenomena in light and sound waves. Furthermore, the phase displacement of the oscillations of the individual radiators may be regulated by tuning. Thus a variety of interference patterns may be created, and analysis of these possibilities shows, that an efficient uni-directional radiation by such methods should be possible.

Methods for unidirectional radiation have been established through the well-known work of Bellini and Tosi. Thru the courtesy of Mr. Bouthillon of the French Post Office, results of tests made in France have been placed at the disposal of the author, which show conclusively directive radiation by the Bellini and Tosi antenna.

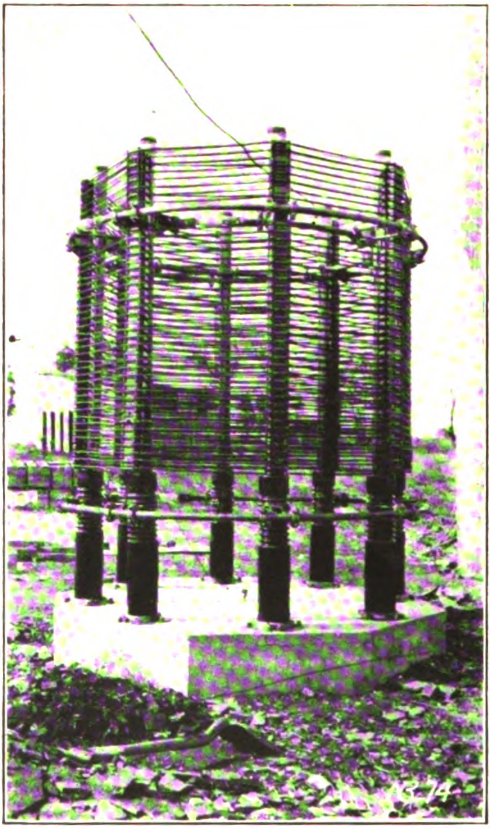


FIGURE 15 -Outdoor Tuning Coil

With the dimensions of antennas used up to the present time, efficient directive radiation has not been practical. It has, however, been proven by various tests that the system of a central power source and a distribution system of energy to a large number of multiple radiators places at our disposal means for constructing radiators of dimensions of one wave length or more.

The New Brunswick antenna (1,500 meters or 5,000 feet long) has a minimum wave length of 8,000 meters as a single antenna, whereas it can be operated as a multiple antenna at 2,000 meters wave length. A detailed analysis of the possibilities of multiple radiation would fall outside the scope of this paper, but the author is in position to predict with confidence that directive radiation on large scale will not only prove practical, but that it will be the most efficient method of radiation.

To add directive radiation to the proposed program for increasing the capacity of radio traffic would perhaps be premature until it has been demonstrated on a large scale. However, it deserves mention in order to show that new principles which may be utilized for still greater expansion of the radio technique have not yet been exhausted.

SUMMARY: The possible limitation in the number of long distance radio stations simultaneously in operation is considered. It is shown how high-speed transmission, improved directional selectivity, and better frequency selectivity would greatly increase the number of feasible co-existent stations,

The utilization of a wave length band as a result of high-speed transmission is discussed in connection with frequency selectivity.

The 200-kilowatt alternator station at New Brunswick is described, special mention being made of the constant speed regulation system for the induction motor drive, the functioning of the magnetic amplifier, and the operation of the multiple antenna. In this latter connection, the possibilities of effective directional radiation are considered.

MEASUREMENT OF THE ELECTROMAGNETIC FIELD OF WAVES RECEIVED DURING TRANS-OCEANIC RADIO TRANSMISSION*

BY

GIANCARLO VALLAURI

(DIRECTOR OF THE ELECTROTECHNICAL AND RADIOTELEGRAPHIC INSTITUTE OF THE ROYAL ITALIAN NAVY, LEGHORN, ITALY)

1

In the early part of August, 1919, the radio station at Annapolis, Maryland, in the United States (call signal "N S S") carried on experimental transmission lasting thru the 24 hours of the day. An occasion thereby arose for the Electrotechnical and Radiotelegraphic Institute of the Royal Navy to carry on quantitative measurements on the electromagnetic field of the arriving waves. The method chosen was a simple one of the substitution type, and was rendered easy of execution by the fact that the antenna at the Institute is a large radio goniometer. This goniometer consists of two entirely similar loop circuits of nearly triangular form, and so placed that one of them makes an angle with the geographic north of 36° or 216° (and therefore approximately NE and SW) and the other an angle of 126° or 306° (and therefore approximately NW and SE). The area enclosed by each antenna is approximately 1,404 square meters (15,100 square feet).

The distance between Annapolis and Leghorn (measured along an arc of a great circle of the earth) is 6,917 kilometers (4,300 miles), and the azimuth of Annapolis relative to Leghorn is 298° 47′.

It follows that the northwest antenna is almost exactly placed for most favorable reception from Annapolis (with an approximate error of $306° - 299° = 7°$). Furthermore, the northeast antenna is in a very favorable position in order not to be influenced by the signals at all.

In connection with the reception, there was used an amplifier with eight vacuum tubes, having six stages of resistance amplifi-

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*Received by the Editor, September 15, 1919.
Translated from the Italian by the Editor.

cation and two stages of telephone transformer amplification. These last two stages were not used during the experiments. Since the inductance of each antenna is 255 microhenrys, and since the amplifier works conveniently with a circuit of much larger inductance, it was found desirable to insert the amplifier into a secondary oscillating circuit, coupled inductively with the antenna circuit.

2

The method of the measurements can be clearly gathered by an examination of the circuits of Figure 1. It consists in obtaining in the NW-SE antenna circuit the actual signal, and then switching over to the NE-SW antenna circuit and inducing in this new circuit (which is identical in all respects with the first except in direction) a locally regulable and measureable electromotive force. The frequency and amplitude of this local electromotive force can be so varied as to render it identical with that induced in the NW-SE antenna by the electro-magnetic wave coming from Annapolis.

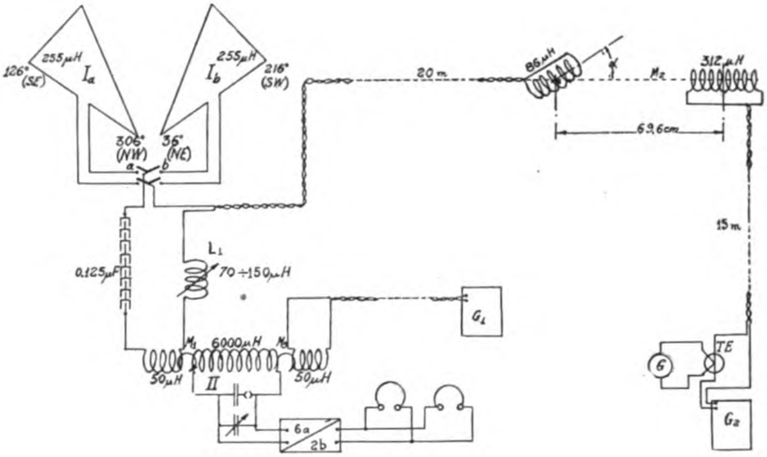

Figure 1—Measuring Circuits

The antenna consists, therefore, of one or other of the two loops (according as the switch *a b* is thrown to one side or the other), of the capacity shown which is made up of eight paper condensers each of one microfarad, connected in series, of the

50-microhenry primary of the mutual inductance M_1 for coupling with the receiver, of the variometer L_1 which can be varied between 70 and 150 microhenrys for tuning, and finally of the secondary of the mutual inductance M_2 for coupling to the local oscillation generator G_2. This last secondary consists of 19 turns of diameter 12 centimeters (4.72 inches), and has an inductance of 86 microhenrys. The secondary circuit II, or receiving circuit, includes the 6,000-microhenry secondary of the mutual inductance M_1, a fixed condenser, and a variable air condenser across which the amplifier is connected. To produce beats in reception, the oscillations produced by the local heterodyne oscillator G_1 are introduced into the receiving circuit thru the mutual inductance M_3. To induce into the antenna circuit, when the switch is thrown to the side *b*, an electromotive force identical with that produced by the Annapolis signals in the NW loop, the oscillations produced by the generator G_2 are introduced into the antenna circuit thru the primary of the mutual inductance M_2. This primary consists of 50 turns of diameter 10 centimeters (3.94 inches), and has an inductance of 312 microhenrys. The current in the circuit of oscillator G_2 is measured by means of a thermo couple TE and a galvanometer G. The calibration of this galvanometer is given in Figure 2 which shows the relation between the reading l of the galvanometer and the current in milli-amperes passing thru the heater wire of the thermo couple. Each of the oscillators G_1 and G_2 is a vacuum tube generator with inductive coupling.

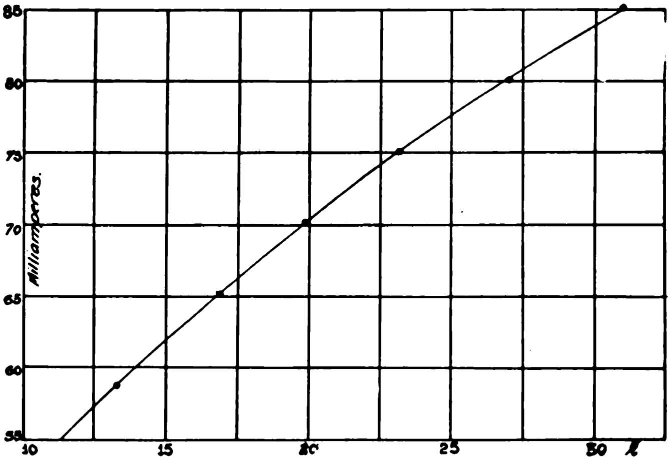

FIGURE 2—Calibration of Galvanometer

Calling α the angle between the axis of the primary and secondary coils of the mutual inductance M_2, it is found that the value of mutual inductance when $\alpha = 0$ (that is, when the primary and secondary inductances are coaxial) may be calculated from Havelock's formula, and is equal to 0.049 microhenrys. In view of the small dimensions of these two coils as compared with their distance apart (69.6 centimeters or 27.5 inches), we can correctly assume that

$$M_2 = 0.049 \cos \alpha$$

in microhenrys.

3

The following is the detailed method of carrying out the measurements. The switch is thrown to side *a* and the oscillator G_1 is started up, while the oscillator G_2 is not in operation. The quantities L_1, M_1, M_3, and the wave length λ_1 of the heterodyne oscillator G_1 are so regulated as to permit convenient reception of the signals. Then, leaving all other constants of the circuits unchanged, one changes over from the *first position* just described to the second position (for which the switch is thrown to side *b*) and the oscillator G_2 is started up. The mutual inductance M_2 is then roughly regulated so as to give approximately the same signal intensity as was obtained in the first position. By close regulation of λ_2, the wave length of the oscillations produced by G_2, there is finally obtained complete identity of the musical tone produced in the two positions. To check up the correctness of adjustment, a number of rapid transfers are made back and forth between the first and second positions. Then, keeping the note constant and leaving λ_2 unchanged, the intensity of the signal is regulated in such a way as to be equal in the two positions. This is accomplished in the second position by varying the angle α on which M_2 depends. When the angle α is 90°, the mutual inductance M_2 is zero and there is always an average sector or region in which the signal produced by G_2 diminishes to the disappearing point. Consequently, there can always be found two angular readings α_1 and α_2 on the two sides of such a sector for which the locally generated signals produced in the second position have the same intensity as the external or incoming signals in the first position. As a matter of fact, for each reading *l* of the galvanometer or rather for each deflection of G_2, four readings of α are taken, two for increasing signal intensity and two for diminishing signal intensity, and the averages of each of the two sets are taken as the values of a_1 and a_2. Then, by passing very rapidly from the second to the first position, it is possible to make sure that the identity of the two signals was really obtained.

4

In the second position (switch thrown to b and oscillator G_2 in operation), if the coupling between circuit I and the oscillator G_2 is exclusively due to M_2, the readings a_1 and a_2 must be symmetrical relative to 90°, or rather the two corresponding values of M_2 must be equal and of opposite sign. As a matter of fact, in spite of the use of screens and the considerable distance of separation between the various parts of the equipment, it is found that the oscillating circuit G_2 does have a certain mutual inductance M_o to the antenna circuit in the second position in addition to the mutual inductance M_2. If we take

$$M_2' = 0.049 \cos a_1$$

and

$$M_2'' = 0.049 \cos a_2$$

as the two values of M_2 corresponding to the two readings a_1 and a_2, we evidently have

$$M = M_2' + M_o = -(M_2'' + M_o)$$

from which

$$M_o = -\frac{M_2' + M_2''}{2}$$

$$M = \frac{M_2' - M_2''}{2} = \frac{0.049}{2}(\cos a_1 - \cos a_2),$$

in which M is the coefficient of total mutual inductance between the circuit G_2 and the circuit I in the second position. That the value M which is that used for the calculation of the induced electromotive force, does not depend on the direct coupling M_o is confirmed by the fact that the results of the measurements, given in the following table and relating to the tests of the same group, are in practical agreement with each other, altho it must be stated that they were attained by an intentional modification of the value of M_o. In order to carry out such modification, the inductance coil of the oscillator G_2 was changed in position, between one measurement and the next, in relation to the remainder of the circuit, or a similar change in position was given to both coils constituting the mutual inductance M_2 The corresponding variations in M_o are indicated by a shifting of the region of silence as indicated on the graduated circular scale of M_2 and particularly by variation of the value of $\frac{a_1+a_2}{2}$.

The results of the measurements would have been invalidated if there had existed a direct coupling between the circuit G_2 and the secondary II, and to avoid any such defect, the two sets of apparatus in question were widely separated from each other. To check up this point, tests were made which showed that no matter how the equipment comprising the mutual inductance M_1 M_3 was placed relative to the remainder of the circuits, no change was produced in the readings a_1 and a_2 for a given reading of the galvanometer and a given signal intensity.

5

There is a danger, however, that the circuit of the heterodyne oscillator G_1 not only influences the secondary II thru the mutual inductance M_3, but also acts on the primary circuit, and that this action may be changed in amount between the first and the second position. As a matter of fact, the coupling M_3 is relatively close, because, in order to secure optimum heterodyne adjustment, it is found convenient to have an electromotive force induced into the secondary of a markedly greater amplitude than that produced by the signal. At any rate, it was possible to prove that such an objectionable influence was actually negligible by taking advantage of the short pause in transmission from Annapolis, and then noting that if the oscillations produced in G_1 and in G_2 remain unchanged and the set is switched from one loop to the other, the readings a_1 and a_2 are indeed somewhat altered, but the difference $(\cos a_1 - \cos a_2)$ remains unchanged.

One may finally consider the effect produced in the second position in the NE loop by the signals from Annapolis. This induced emf. is proportional to $\cos \beta' = \cos\ (299° - 216°) = \cos 83° = 0.12$, and it should be added to that induced from the oscillator G_2. It appears, however, more reasonable not to include this effect, since it is found that in the second position a slight variation in λ_2 does not cause any alteration in the note and since, in the same position, and with G_2 extinguished, even the greatest amplification obtainable does not permit one to hear any trace of the signals from Annapolis.

6

From the reading I of the galvanometer and a_1 and a_2 and the calibration of M_2, the effective value of the induced electromotive force can be calculated from:

$$E = \omega M I = 2.67\ (10)^{-3}\ (\cos a_1 - \cos a_2)\ I$$

where $\lambda = 17{,}300$ m. $\omega = 109.000$

$$M = 49\,(10)^{-9}\,\frac{\cos\alpha_1 - \cos\alpha_2}{2}.$$

It is worth while mentioning that the identity of the note obtained from the signal of Annapolis in the first position and the signal from oscillator G_2 in the second position can also be secured at a frequency f_2 differing from f. which is that of the signal, provided that the frequency f_1 of G_1 satisfies the condition $f - f_1 = f_1 - f_2$. This can be readily avoided by noticing which way the note changes as λ_1 is changed in the first position and also when λ_2 is changed in the second position. In all these measurements, we arrange that $\lambda_1 < \lambda$ and $\lambda_2 > \lambda_1$, so as to have $\lambda = \lambda_2$.

Having obtained the effective value E of the electromotive force induced, we can obtain the magnetic field strength H of the incoming wave from Annapolis from the following equation:

$$H = \frac{E\,(10)^8}{\omega\,S\cos\beta} = 6.58\,(10)^{-5}\,E,$$

where S equals 14.040,000 square centimeters (15,100 square feet), this being the area covered by the receiving loop, and $\beta = 7°$ is the angle between this plane and the horizontal direction of propagation of the incoming wave. The effective value H of the magnetic field is expressed as C. G. S. electromagnetic units. From this, there can be obtained the electric field intensity $F = vH$ where v is the velocity of light, and therefore

$$F = 3\,(10)^7 \cdot H \text{ volts per km.} = 1.97\,(10)^3 \cdot E \text{ volts per km.}$$

Carrying out the calculations in the manner indicated above, the following numerical table of result has been obtained. The observations were always carried out with two observers, each using one of the headband telephone receivers provided with the amplifier. Six different operators took part in these measurements. In the table given, *all* the results of the measurements carried out are given without any rejection whatsoever. The values may, therefore, be taken as a fair representation of the accuracy of the results from the point of view of accidental errors. As can be easily seen, with the method of measurement chosen it is by far preferable to take readings when the transmitting station makes a long dash rather than when it is transmitting signals.

Observation	Group	Date	Time G.M.T.	α_1	α_2	$\cos\alpha_1 - \cos\alpha_2$	l Galvanom.	I_2 mA	E μV		H C.G.S.	F Volt/Km.	hI Amp. x Km. (Austin)	(Fuller)	Signals from NSS
1	1	Aug. 6, 1919	14h 6m	71° 0	94° 0	0.3954	26.0	78.7	83	77	$5.0.10^{-9}$	0.151	530	78	Long Dash
2			9	73° 5	93° 5	3450	27.2	80.3	74						"
3			12	71° 0	91° 5	3518	26.5	79.4	75						"
4	2	"	21h 38m	70° 5	93° 0	0.3861	14.0	60.0	62	53	$3.5.10^{-9}$	0.104	360	82	"
5			42	69° 0	88° 0	3235	13.0	58.2	50						"
6			47	70° 0	88° 0	3071	12.5	57.2	47						"
7			51	69° 0	88° 0	3235	14.0	60.0	52						"
8			57	69° 0	90° 0	3584	12.5	57.2	55						"
9	3	Aug. 7, 1919	5h 50m	74° 5	84° 0	0.1627	18.0	67.0	29	34	$2.2.10^{-9}$	0.067	230	33	Signaling
10			59	91° 0	105° 5	2497	13.0	58.2	39						"
11			6h 3m	94° 0	106° 0	2058	16.0	63.6	35						"
12			6	94° 0	104° 0	1721	16.0	63.6	29						"
13			8	96° 5	107° 0	1792	17.5	66.2	32						Long Dash
14			11	93° 0	107° 0	2401	16.9	65.0	42						"
15	4	"	7h 24m	94° 0	102° 0	0.1381	17.5	66.2	24	25	$1.6.10^{-9}$	0.049	170	24	"
16			26	95° 5	101° 5	1036	17.8	66.6	18						"
17			28	93° 0	105° 5	2149	12.3	56.8	33						"
18			30	93° 5	102° 0	1469	15.0	61.8	24						"
19	5	"	9h 8m	91° 5	101° 0	0.1646	25.0	77.5	34	36	$2.4.10^{-9}$	0.071	250	38	"
20			10	73° 2	86° 5	2280	18.8	68.4	42						"
21			13	74° 0	84° 0	1711	21.0	71.9	33						"
22			16	74° 0	84° 0	1711	22.0	73.4	34						"
23	6	"	11h 51m	72° 5	89° 5	0.2920	13.0	58.2	46	42	$2.7.10^{-9}$	0.082	280	41	"
24			53	73° 5	87° 7	2439	12.0	56.3	37						"
25			57	73° 2	88° 0	2551	13.7	59.5	41						"
26			12h 1m	88° 5	104° 0	2681	13.8	59.7	43						"
27			4	73° 5	89° 2	2700	14.1	60.2	43						"
28	7	"	16h 23m	73° 7	85° 5	0.2021	17.6	66.3	36	39	$2.5.10^{-9}$	0.076	260	38	"
29			26	72° 5	85° 0	2135	16.1	63.7	36						"
30			30	71° 7	88° 0	2791	11.3	54.8	41						"
31			32	71° 0	88° 0	2907	12.7	57.5	45						"

It is easy to understand that the method might have been equally well carried out, using appropriate precautions, by the employment in the second position of an artificial antenna equivalent to, and in the place of, the NE-SW loop. However in this case the results of the comparison would have been less certain because atmospheric disturbances might have been missing in the second position. During the days on which the tests were carried on, these disturbances were very intense and did not appear to change on switching over from one antenna to the other.

7

The results of the measurements are not sufficiently numerous to permit drawing definite conclusions regarding the daily variation in the intensity of the signals received. For this purpose it would also be necessary to know the actual variations in antenna current at Annapolis during the experiments. The measurements of group 4 seem to confirm the "fading effect" of signals received from America during the morning hours, an effect which has been noticed at all the Italian stations which receive from North America.

The results obtained supply the quantitative material necessary either for verifying the various formulas enabling calculation of the range of a radio station or for providing the values of the minimum electromagnetic field intensity necessary with a given type of antenna to enable satisfactory reception. In this connection it may be stated that with an induced electromotive force of about forty microvolts, and therefore with a magnetic field of $2.6\ (10)^{-9}$ C. G. S. units and an electric field of 0.078 volts per kilometer, reception of messages using a eight-step amplifier was readily accomplished except during the periods of maximum atmospheric disturbance. As an example, during the afternoon of August 5th, the messages sent from Annapolis could be copied with perfect continuity at any time and were received with an intensity of six. (In this system, 4 = weak, 5 = somewhat weak, 6 = fair, 7 = good, 8 = strong, 8 = very strong. The normal signal strength is 6.)

Considering the formulas dealing with the calculation of the range of radio telegraphic stations, and particularly that of Austin* which can be put into the following form:

$$I_r = 377 \frac{h_r\, h\, I}{\lambda\, d\, R} \sqrt{\frac{\theta}{\sin \theta}}\, \varepsilon^{-0.0015 \frac{d}{\sqrt{\lambda}}}$$

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* PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, February, 1917 volume 5, number 1, page 25

we have

$$F = \frac{R I_r}{h_r} = 377 \frac{h I}{\lambda d} \sqrt{\frac{\theta}{\sin\theta}}\, \varepsilon^{-0.0015 \frac{d}{\sqrt{\lambda}}}$$

where R is the resistance of the receiving antenna, h_r its effective height, I_r the current intensity at its base, h and I the corresponding effective height and current in the transmitting antenna, λ the wave length used, d the distance along a great circle between the two stations, and θ the angle subtended at the center of the earth by the arc joining the two stations. R is expressed in ohms, I and I_r in amperes, h, h_r, λ and d in kilometers, while the electric field F is expressed in volts per kilometer.

Starting from the unknown quantity hI and taking $\lambda = 17.3$, $d = 6{,}917$, $\theta = 62° \, 18.5'$, we have

$$h I = \frac{\lambda d}{377} F \sqrt{\frac{\sin\theta}{\theta}}\, \varepsilon^{0.0015 \frac{d}{\sqrt{\lambda}}} = 3{,}500\, F \quad \text{amp. km.}$$

The values of hI which are obtained in this way from Austin's formula, and which are given in the table, are evidently too large and actually about ten times larger than the true values. This indicates that for long distances Austin's formula gives values of the intensity of the electromagnetic field of the incoming wave which are much smaller than those actually found, or, in other words, that propagation of the waves is accomplished in a more favorable way than that indicated by the assumption on which the formula is based.

Very similar results are obtained by applying Dr. L. F. Fuller's formula ("Proc. American Institute of Electrical Engineers," volume 34, page 567, 1915):

$$I_r = 377 \frac{h_r h I}{\lambda d R} \sqrt{\frac{\theta}{\sin\theta}}\, \varepsilon^{-0.0045 \frac{d}{\lambda^{1.40}}},$$

whence

$$h I = \frac{\lambda d}{377} F \sqrt{\frac{\sin\theta}{\theta}}\, \varepsilon^{-0.0045 \frac{d}{\lambda^{1.40}}},$$

or, in this case, $hI = 500\, F$ approximately.

The values of hI thus obtained are about one-seventh as great as those given in the table. Knowing the effective values of hI for each group of observations in the table, there can be calculated for each of them the exponential coefficient to be substituted for 0.0045 in Fuller's formula to obtain agreement between the experimental results and the formula.†

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† The average value of the antenna current at Annapolis for the hours during which the seven groups of tests were carried on was 210 amperes. Taking the effective antenna height as 105 m. (320 feet), we have for hI the value of 22.1 amp. km. These values have been obtained by courtesy of the United States Navy Department thru Captain S. W. Bryant, Acting Director of Naval Communications.

It is hoped that the opportunity will arise to carry out much more numerous and definite measurements of this type which will require the use of several different stations, at different distances from each other, and employing various wave lengths.

Leghorn, August 22, 1919.

SUMMARY: A series of recent experiments is described on the determination of the electric and magnetic field intensities of the waves received at Leghorn, Italy, from the Annapolis, Maryland transmitting station.

A pair of large loops, respectively practically parallel and perpendicular to the incoming wave are used. The incoming signals are received by heterodyne oscillator and amplified. The substitution method is used for the measurements, whereby the incoming signal is compared to an artificial signal produced in the non-receptive loop by a local oscillator. A thermo-couple and galvanometer in the oscillator circuit, together with a calibrated mutual inductance for coupling to the non-receptive antenna, permit quantitative results. The necessary precautions and procedure are described.

Numerical results are given, and it is shown that the field strengths of the incoming waves are much greater than those calculated by the usual transmission formulas.

Louis W. Austin (by letter): I am very glad that so eminent an experimenter as Professor Vallauri has entered the field of quantitative measurements in long distance transmission, in which I have been left so long practically to myself. For the further progress of transmission theory it is imperative that a large amount of quantitative data be gathered in different parts of the world, especially on signals coming from great distances. It has become increasingly plain that no single formula can represent the experimental facts correctly at all times. There are factors active in transmission which produce such variations in signals travelling more than two thousand miles (3,200 km.), that on certain days the signals produce current in the receiving antenna in Washington more than fifty times the strength produced on the weakest days. This is quite apart from all experimental errors and from the special fading periods lasting a few hours. It is possible that this region is particularly unfortunate in the matter of variations and very probably in other parts of the world the signals may be found more constant, but it is certain that the best that can be expected from a transmission formula is that it shall represent a fair average of the strength of the received signals throughout the year.

While Professor Vallauri's method of measurement is rather complicated, all the observations and necessary precautions appear to have been taken with the greatest care. The only improvement which I can suggest would be a further check on the calibration of the receiving system by sending with a small known antenna current from a portable antenna or loop at a distance of one or two wave lengths from the receiving station. At this distance and with the long wave length used, absorption would not be appreciable, and all the constants and the distance being known, the original calibration could be verified with considerable certainty.

From the several thousand long distance measurements which have been made by the Laboratory in Washington and at other points, I feel sure that the Navy formula will agree with the *average current for the year* received at Leghorn from Annapolis with an error not much greater than fifty per cent.

It is now believed that the attenuation over salt water of damped and undamped signals is not appreciably different, as the formula originally developed from spark observations seems to apply to undamped measurements.

Fuller's formula for undamped waves ("Transactions of the American Institute of Electrical Engineers," volume 34, page 809, 1915), which gives a smaller attenuation than the Navy formula for long waves is probably in error on account of the assumption in the experimental work that the received telephone current, using a slipping contact tikker, is proportional to the power in the antenna, while, as a matter of fact, it is more nearly proportional to the received antenna current. At any rate it has not been found possible to make this formula represent the observations made in the Research Laboratory, altho at certain times it does appear that the longer wave lengths have a greater superiority over the shorter than would be represented by the Navy Formula.

RADIO DIRECTION CHANGES AND VARIATIONS OF AUDIBILITY*

By

Carl Kinsley

(Major, Signal Corps, United States Army)

and

Albert Sobey

(Major, Signal Corps, United States Army)

Introduction

The development of methods for the determination of the point from which radio signals originate was actively carried on during 1918. The radiogoniometer measurements played an increasingly important role during the fall of last year. It seemed probable that the so-called radio compass might be called on to an extraordinary extent to assist in the projected operations of the spring of 1919.

There has been no opportunity, however, for an examination of the degree to which reliance could be placed upon that method of determining directions when the origin of the radio signals was known. During the winter and spring of 1919, we were fortunately able to carry on an extended series of measurements under the peculiarly favorable circumstances of widely separated stations operated by well-trained observers.

Discussion of Problem

There has been an extensive examination of the possible means by which the electromagnetic wave is bent around the earth without suffering the attenuation which would be present if the transmitting medium were homogeneous. The wave apparently follows the surface of the earth with but little of its energy being dissipated by a vertical expansion of the wave. This may be due to there being an upper permanently ionized region known as the Heaviside layer at an altitude of about 80

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* Received by the Editor, August 5, 1919.

km. (50 miles), the lever at which the aurora is first observed. which acts as a conductor so that the wave spreads between it and the earth as tho they were two concentric conducting spheres with a dielectric between.

If this should not be the case, then the wave might be refracted around the earth without getting far from the earth's surface. Several possible causes for such a refraction have been suggested. Eccles shows how this might come about by slowly increasing ionization and Schwers considers that the hygrometric state may be such that the electromagnetic wave will be totally refracted downward and follow the curvature of the earth.

In any case, the wave passes thru regions where the moisture content is considerable, and where the ionization is partially dependent on the sun's light. Since the transmitting medium is not homogeneous, it is obviously possible that there should be some horizontal bending of the waves which might appreciably effect the radiogoniometric measurements.

It is desirable, therefore, to determine to what extent the electromagnetic waves, used for radio purposes, change their direction as measured by the radiogoniometer, and to obtain, if possible, the variations in audibility of the signals so as to note the relation existing, if any, between changes of audibility and changes of direction.

The causes for departure from the direct line of the transmission of radio signals as well as for their constantly changing audibility, may be expected to be explainable on a basis of changes in the uniformity of the transmitting medium. This will be considered at some length for the purpose of determining whether or not the physical conditions existing afford the necessary discontinuities to explain the variations observed.

METHODS AND RESULTS OF RADIO TESTS

In the measurements given in this paper, accurate quantitative methods were employed in the direction measurements.

Having in mind the different methods used by the English and French radio engineers in direction measurements, the following brief statement is given as it is believed that the method employed renders any considerable error impossible, altho the accuracy is much greater than that found to be necessary after the establishment of the fundamental facts of radio direction changes have been made.

Figure 1 gives a polar curve of audibility changes as the receiving loop is rotated. This curve is thoroly characteristic.

The difference between its upper and lower parts is due to the design of the receiving loop, which was purposely constructed so as to indicate the absolute direction of the radio signals. It has been found from many tests that the minimums lie 90 degrees from the direction of the transmitting station—using normal daylight conditions. It is to be noted that this is true even tho the maximum vector does not lie along this direction. In the figure shown, a difference of about 10 degrees will be noted while the direction of the transmitting station as computed from the minimum values is in error by only 35 minutes.

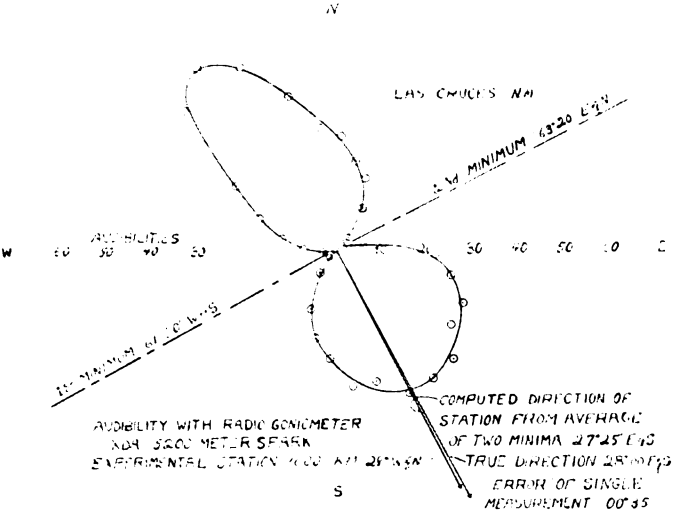

Figure 1

Figure 2 illustrates the method employed in determining the exact position of two of the minimums from which the directions plotted in Figure 4 are obtained. The particular points chosen were those showing the largest variations from the correct direction observed during the week of the test. These particular values were taken March 19th, upon which date no rain fell at Houlton, but at every other town listed in Maine. It will be noted that each minimum is obtained as the result of nine readings of audibility from which are plotted a medial line which intersects the curve of audibility at its correct minimum. It will be seen that an error larger than ¼ degree would be unlikely. Other illustrations are given in Figure 7.

The audibility measurements thus need only be relative in order to give an exact quantative determination of the location of a minimum. We employed, therefore, the method of measurement involving the use of an audibility meter having an iron core inductive resistance in series with a high resistance telephone—2,000 ohms shunted by non-inductive resistances, the audibilities being computed by using the impedance of the telephone, not its resistance, in the usual equation. It was essential that the audibility measurements be made quickly so that an appreciable change of the direction would not occur during the time needed for making the nine readings. This time varied from three to seven minutes. If a change in minimum took place then the curve plotted was unsymmetrical and the medial line might be far from vertical. In the illustrations shown, the change in direction was of unusual rapidity, but the measurements were still sufficiently rapid to establish with certainty the position of the minimum. When the changes are too rapid to allow the use of the accurate method outlined then a method of disappearing points is used as illustrated in Figure 8.

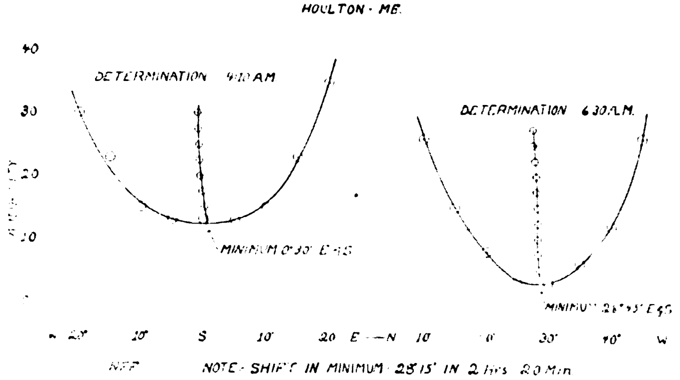

FIGURE 2

In Figures 3, 4, 5, and 6, the direction determinations are plotted radially and the time as the angles so that if the direction of the radio signals had not varied the plot would have shown merely a true circle.

The loop used at Houlton, Maine, for these measurements is as follows: sixty turns, number 26, B. and S. solid copper wire,* double cotton-covered, wound in one plane with horizontal wires spaced 0.4 inch (1 cm.), and vertical wires 0.6 inch (1.5 cm.) apart. The outside dimensions of the loop are therefore, vertical dimensions 6 feet 5 inches (1.95 m.), and horizontal dimensions 9 feet 10 inches (3 m.). The loop was 7.5 inches (19 cm.) below ceiling and 9 inches (22.9 cm.) above floor, and having clearance of 6 inches (15.2 cm.) on one side and 12 inches (20.5 cm.) on the other, in a room on the lower floor of a small detached wooden house. The vertical axis about which the loop was turned was pivoted to floor and ceiling.

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* Diameter of number 26 wire = 0.06 inch = 0.15 cm.

Figure 3 gives a series of direction determinations made as described above and each direction is the result of a computation from the accurately determined two minimums of a polar curve, each of which required nine audibility readings.

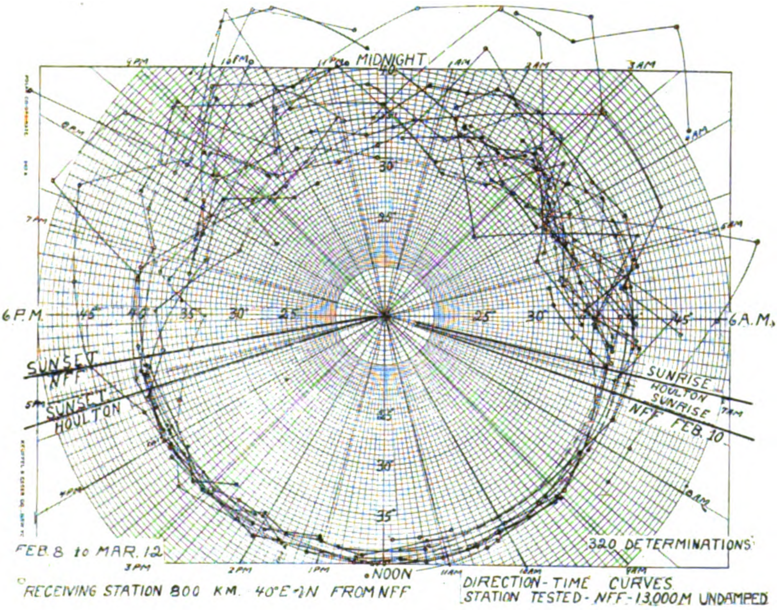
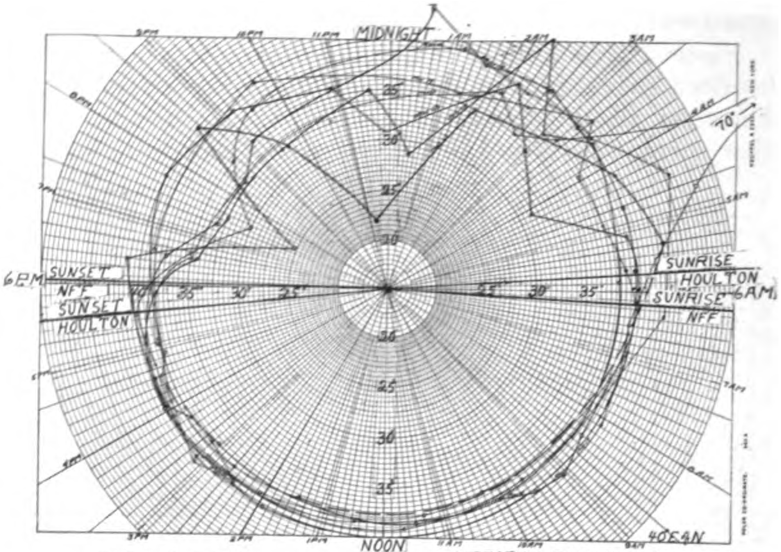


FIGURE 3

It will be noted that, during the sunlight hours, the direction measurements varied but little and the values could be used for determining the location of the transmitting station. During the night, however, the direction measurements showed extremely erratic values.

Figure 4 was obtained by the same methods as in Figure 3, and all the measurements made during one week of continuous observation are shown. The particular case exhibiting the widest variation found during this week—4.10 A.M. to 6.30 A.M. in March 19th—has included in it the original data, a part of which is given in Figure 2 to show the accuracy of the measurements.



MARCH 12-19-1919.

DIRECTION-TIME CURVES
STATION TESTED-NFF-13,500 M. ARC.
RECEIVING STATION 800 KM. FROM NFF.

FIGURE 4

Figure 5 was obtained in the same way as in the case of the preceding figures, and it is to be noted that in the transmitting station, which also employed a machine generator, was located 12,000 kilometers (7,500 miles) from the experimental station, and most of that distance was over water. During the time when the whole path was in sunlight, the measurements showed no sudden changes in direction, altho there were a considerable number of observations taken on both sides of the mean.

Figure 6 is designed to show the directional measurements taken at widely separated places using radio transmission from the same station, which was a low frequency spark of 5,700 meters wave length. At McAllen, Texas, the receiving loop was of approximately similar design, but only thirty turns were employed. It is to be noted that the transmitting station was

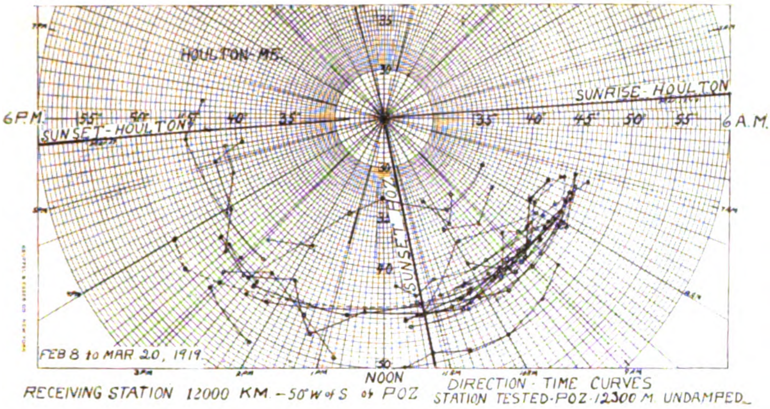


FIGURE 5

900 kilometers (560 miles) from one of the experimental stations, while in the case of Figures 3 and 4, the distance apart of the transmitting and receiving stations was approximately the same,

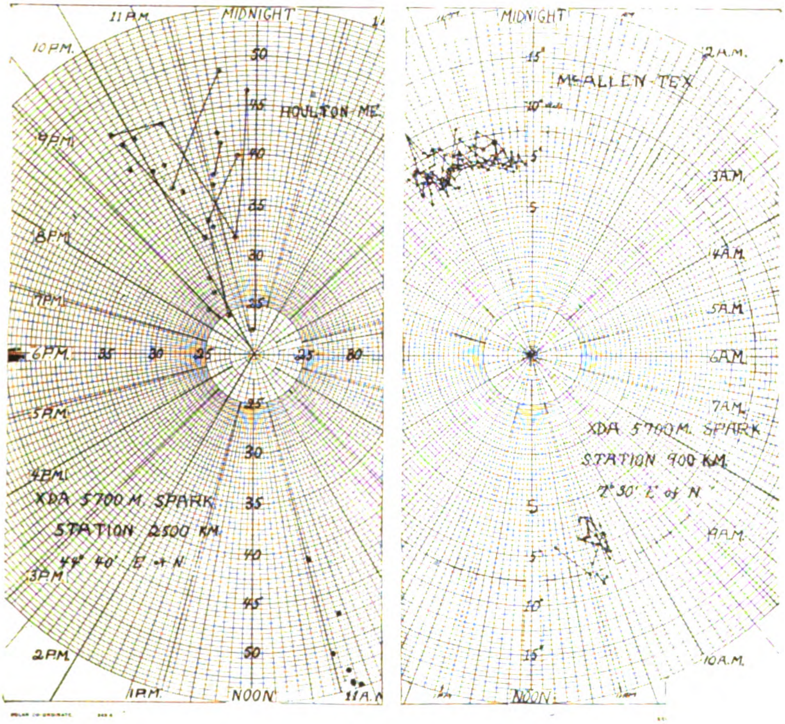


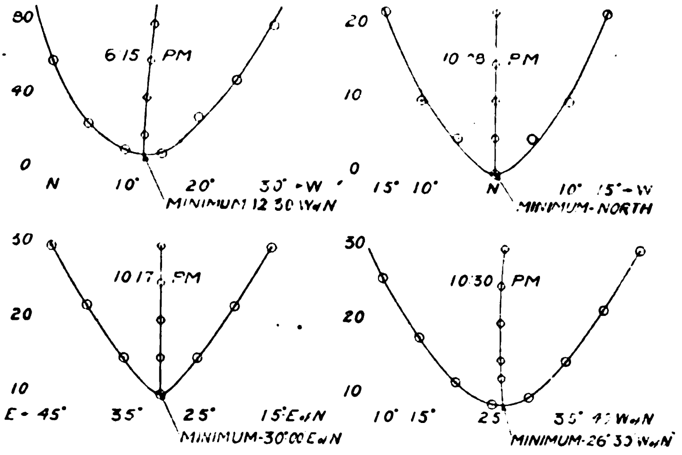
FIGURE 6

810 kilometers (406 miles). A more distant station, Houlton, Maine, 2,500 kilometers (1,560 miles) from the transmitting station, exhibits wide variations in direction during the night hours and considerable variation during the day time. The transmitting station is only operated for a limited time during the twenty-four hours, so no determination could be made as to whether or not there was a sharply defined change in the condition at sunrise, and sunset. It is not believed, however, that the large directional changes are due to the greater distances but rather to conditions relatively local.

It is to be noted that Figure 1 shows a complete audibility curve for this station, and the measurements obtained in that case indicate a true direction for the transmitting station. It will be seen from the observations taken at McAllen, Texas, that in general it was found at that point that the direction measurements seldom varied materially from the true direction of the station.

Figure 7 gives four of the minimums obtained at Houlton, Maine, in determining the variation of the direction of the radio signals sent by the Arlington, 2,500 meters wave length spark, on March 29th.

A considerable number of determinations were made on this station, and it was found that the direction very frequently differed from the true direction of the station.



NAA 2500M SPARK — NOTE: SHIFT IN MINIMUM 56° 30 IN 13 MINUTES

FIGURE 7

The complete measurements always involved the determination of both of the minimum curves, but only one of each pair is given in the illustration. It will be found that there can be no question as to the accuracy of the measurements, and the reality of the tremendous change in the measured direction. In spite of the large and rapid change, the method employed was sufficiently fast to insure a symmetry of the curves, and consequently the determination of the direction is reliable.

Figure 8 illustrates a curious change observed on March 2nd in the case of the same transmitting station used to obtain the results shown in Figures 3 and 4. There was a slight rain at Houlton on that day. In the course of these measurements, it was found that the direction was changing so rapidly that the audibility curve showing the minimum was no longer symmetrical, and consequently it became necessary to use a more rapid method of determining direction. The different methods of observation used are indicated in the drawing.

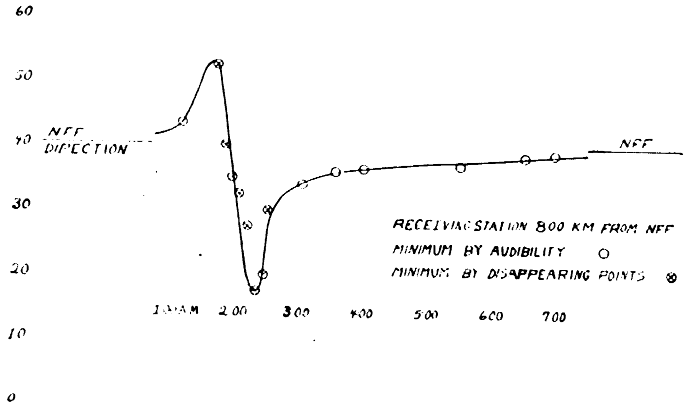


FIGURE 8

The rapid measurement of direction was accomplished with sufficient accuracy for most purposes by rotating the loop until the signal fades out or disappears. With a proper maximum audibility, it is generally possible to obtain, therefore, two pairs of disappearing points, giving an average from which the position of the loop for minimum audibility is determined. This method leaves a good deal to the judgment of the operator, and so one is inclined to place less reliance upon this method except for the particular case when he himself is the experimenter. It is believed, however, that good relative values may be obtained in this way, and that the results reported by any one operator would give reliable information with regard to the changes in audibility.

In the remaining figures are shown audibility measurements and directional measurements, in which the relative direction only was desired in order to determine whether or not the changes in audibility and the changes in direction have any relation to each other.

The directional measurements were therefore made by the disappearing point method, which has a very considerable advantage for this type of experiment, in that the time needed for making the directional measurements is only about 30 seconds, and so it is possible to obtain very sudden changes of direction, with greater relative accuracy.

Figure 9 illustrates the phenomenon often observed, and known as "swinging" of the audibility. The audibility measurements were made by the shunted telephone method, and so varied by steps of considerable magnitude. This is the cause of so many of the observations showing exactly the same audibility while a more accurate method would have distributed the values.

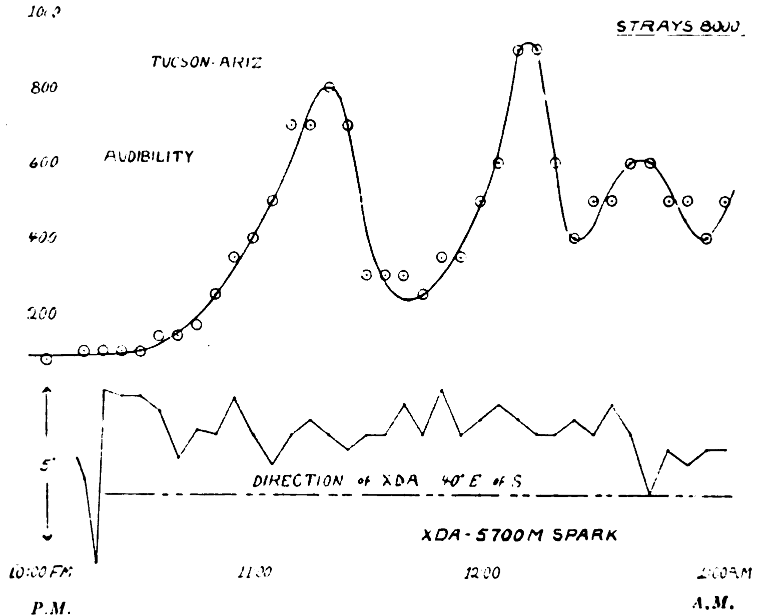


FIGURE 9

During the time covered by these observations, there was but little change in the direction measurements, and the only rapid change in direction noted occurred during the time that the audibility of the signal was constant. These observations were taken June 16th with heavy strays but with a clear sky.

It should be stated that the audibility measurements were made not with the use of the loop, but instead by using a vertical antenna which was located at more than 0.2 kilometer (610 feet) from the loop station; and it was proved by experiment that no interference existed between the two receiving stations, independently employed in making audibility measurements, and in determining by the loop the direction of the radio transmission.

In the following figures the same methods outlined above for determining audibility of the signals and the directional measurements, were employed.

Figure 10 is primarily intended to show that the same station may exhibit widely fluctuating values of audibility with negligible changes in direction, and also the converse of large and rapid changes in direction with inconsiderable changes in audibility. The data for these curves were obtained on June 4th. The weather was reported as hot, with the day partly overcast.

There are shown also in this figure values for strays which

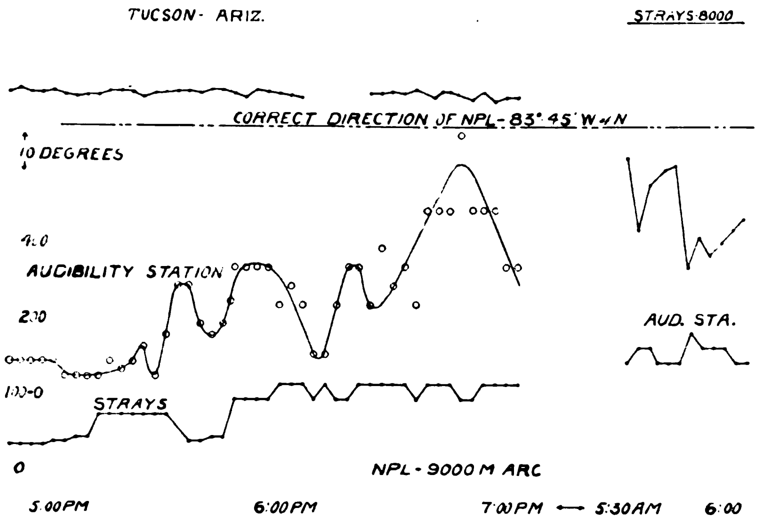


FIGURE 10

in the first case were comparatively low, and in the second were extraordinarily high. It is believed, however, that the presence of high strays is no indication of excessive changes in audibility. This will be brought out in the following figure.

Figure 11 contrasts results obtained from two spark stations, namely, "XDA" using 5,800-meter wave, and "XAJ," using 960-meter wave. It is to be noted that a very considerable change in direction shown by "XDA" is accompanied by a large decrease in strays, but that there was little change in the audibility of the station.

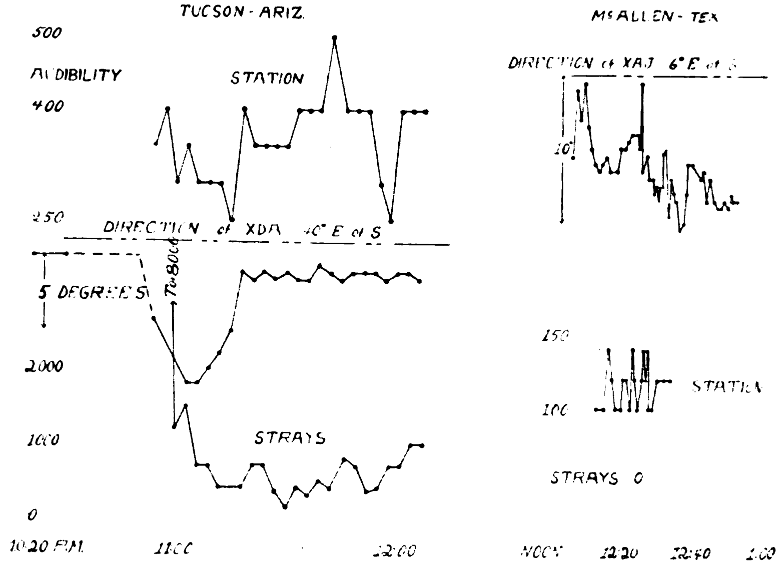


FIGURE 11

The results obtained from XAJ were taken between noon and one P.M., and exhibits much larger changes in direction than are normally found during the day time. It should be noted that, in this case, strays were absent and the audibility of the station was substantially constant.

The observations taken on XDA were made on May 22nd, at Tucson, Arizona. There was during the day time a rainfall of 0.34 of an inch (0.7 cm.). The station observations taken on XAJ were made upon the same date, May 22nd, at McAllen, Texas. There was no rainfall at that point.

On the same day, at McAllen, observations were taken on XDA, which showed wide variations in direction, that were displaced by about one hour and a half from those observed at Tucson. The exact observations are not given, however, since a certain difficulty with the apparatus make the results less reliable than any of those reported in this paper. Simultaneous readings have been made at the three widely separated stations of Houlton, Maine, Tucson, Arizona, and McAllen, Texas, but the results are not yet ready for publication.

Figure 12 illustrates a very special case where the radiogoniometer loop exhibited no directional effect when rotated above a vertical axis. At the same time, on July 11th, the same transmitting station was observed at both Tucson, Arizona, and Houlton, Maine, and thruout the whole period at Tucson no direction could be obtained. The weather condition at Houlton, Maine, was overcast sky but no rain, while at Tucson, Arizona, there were bad static conditions, with lightning around the horizon but no precipitation of rain.

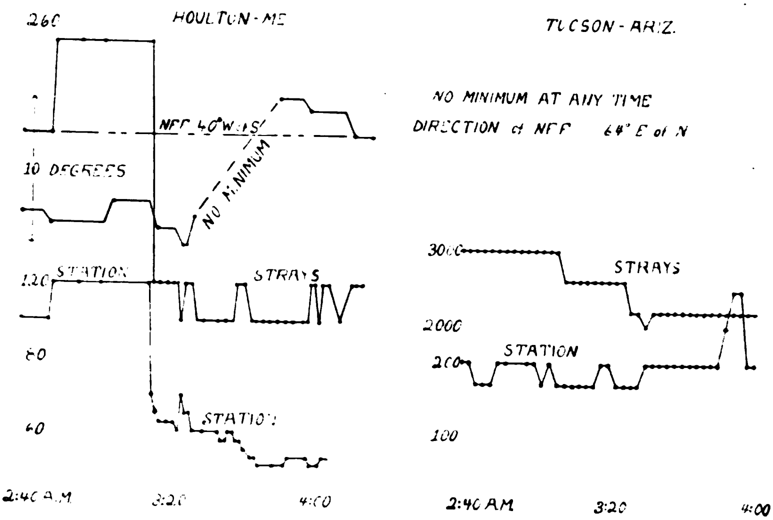


FIGURE 12

The observation at Houlton, Maine, showed uniform direction measurements somewhat displaced from normal, and with a broadening of the minimum. This means that the changes of audibility in the neighborhood of the minimum position of the loop, as the loop is rotating, were less rapid than normal and also with a maximum audibility of the usual value it would be found that the signals did not completely disappear when the loop was in the condition showing the minimum audibility; frequently it is necessary to determine the position of the loop for minimum by a single observation under these special conditions, since there is no complete disappearance of the signal as the loop approaches its position for minimum audibility.

When the special condition of "no minimum" is observed, the loop may be rotated about its vertical axis without any appreciable change in audubility. The explanation of this condition is given in a subsequent paragraph.

TABLES

No minimum audibility as radiogoniometer loop was rotated about a vertical axis.

NUMBER 1

March 6th and 7th, 1919. Houlton, Maine.

1 A.M. to 7 A.M., March 7th, 1919, Sergeant Heinline.
1.15-1.28 A.M. NFF 36 degrees 90 minutes E. of S.
2.00-2.10 A.M.

Note.—Took readings on NFF and NSS and plotted curves for same. While reading NFF from 2 A.M. could find no minimum point; audibility being practically level at all points of the compass. Readings taken on NSS immediately after showed normal conditions for that station, while NFF continued flat.

2.10-2.24 A.M. NSS 42 degrees 30 minutes.
3.12-3.25 A.M. NFF 27 degrees 00 minutes.
3.27-3.36 A.M. NSS 43 degrees 45 minutes.
4.45-4.55 A.M. NFF 41 degrees 45 minutes.

Weather conditions were unsettled, March 6th, rain 0.20 inch (0.5 cm.); March 7th, rain 0.10 inch (0.25 cm.).

NFF is New Brunswick, New Jersey, employing a machine transmitter giving undamped waves at 13,500 meters.

NSS is Annapolis, Maryland, employing a large arc transmitter giving undamped waves of various lengths, and reported to have been 15,000 meters at this time.

NUMBER 2

6 P.M., April 24th to 1 A.M., April 25th, 1919.

Sergeants Pfeiffer and McAllister

Glace Bay, "GB," with 8,000 *meter spark*

10.24 P.M. 12 degrees 30 minutes.
11.02 P.M. 12 degrees 30 minutes.
10.43 P.M. 12 degrees 30 minutes.
11.20 P.M. 12 degrees 30 minutes.

Arlington, "NAA," with 2,500 *meter spark*

11.30 P.M. *Note.*—NAA was tuned in on both of the loop receiving sets, but in neither case was it found possible to get definite disappearing points.

Mexico City, "XDA," with 5,700 *meter spark*

11.38 P.M. 27 degrees 30 minutes W of N.
11.59 P.M. 25 degrees 0 minutes.
11.49 P.M. 27 degrees 30 minutes W of N.
12.01 A.M. 42 degrees 30 minutes.
11.50 P.M. 29 degrees 0 minutes W of N.
12.25 A.M. No minimum.

Note.—Disappearing points began to spread, and in next minute no minimum points could be distinguished. This condition continued until 12.32 A.M., when it quickly changed and the following readings were taken.

12.34 A.M. 26 degrees 0 minutes.
12.52 A.M. 26 degrees 0 minutes W of N.

Weather conditions were unsettled, and traces of rain fell on the 24th and 25th about one-third of an inch.

Note.—In both number 1 and number 2 it is of importance to note that while certain stations gave *no minimum* reading when the radiogoniometer loop was rotated about its vertical axis, other stations gave good direction measurements.

NUMBER 3

July 11th, 1919, 1 to 8 A.M. Houlton, Maine.

Master Signal Electrician Pfeiffer

New Brunswick, New Jersey, "NFF," with 13,500 *meter undamped wave*

3.05 A.M. 35 degrees 30 minutes.
3.24 A.M. 32 degrees 30 minimum points much broader.
3.17 A.M. 33 degrees 45 minutes.
3.27 A.M. 34 degrees 30 minutes minimum points broad.

Note.—From 3.33 A.M. to 3.58 A.M., the minimum points were too broad for reading.

3.50 A.M. 42 degrees 30 minutes minimum points sharp.
3.59 A.M. 41 degrees 45 minutes normal.

Weather, overcast—light south breeze, cool, no rain.

NUMBER 4

July 11th, 1919, 8 P.M. to 4 P.M. Tucson, Arizona.

Corporal Wm. F. Aufenanger

New Brunswick, New Jersey, "NFF," with 13,500 *meter undamped wave*

Note.—NFF heard from about 3.00 A.M. to end of watch, but was very weak. During transmitting periods neither Captain Ives nor operator could discern any disappearing points, as signals were faintly audible over entire 360 degrees. Weather and static conditions unusually bad, lightning continuously playing around horizon.

NUMBER 5

July 13th, 1919, 1 to 8 A.M. Houlton, Maine.

Master Signal Electrician Pfeiffer

New Brunswick, New Jersey, "NFF," with 13,500 *meter undamped wave*

3.20 A.M. 43 degrees 30 minutes broad minimum.
4.02 A.M. 41 degrees 0 minutes sharper minimum.
3.57 A.M. 42 degrees 0 minutes broad minimum.
4.07 A.M. 44 degrees 0 minutes broad minimum.
4.29 A.M. No reading. *Note.*—Minimum points too broad reading.

4.39 A.M. No reading. *Note.*—Minimum points hardly discernible, apparently being around due east and west.

4.47 A.M. No reading. *Note.*—Minimum points continually changing but very indefinite, now being around due north and south.

4.49 A.M. 55 degrees 0 minutes broad minimum.

4.57 A.M. 44 degrees 30 minutes sharp minimum.

4.52 A.M. 48 degrees 0 minutes slightly sharper.

Weather: Clear, cool, full moon.

It is particularly noted that all of the cases observed when no directional measurements could be made appeared during the night, and also in the large majority of them bad weather conditions prevailed.

DISCUSSION OF RESULTS

It has been shown by the preceding figures that radio telegraphic signals of all kinds, including transmitters employing radio frequency alternators and arcs, giving undamped waves of 4,900 to 17,300 meters length, and various spark systems having damped waves from 960 to 5,700 meters wave length, suffer distortions whether transmission is over land or over water. Many observations have been taken which are not shown in the figures, covering a distance range from 65 kilometers (40 miles). (Annapolis, 17,300-meter arc received at Washington) to 12,000 kilometers (7,500 miles). (Funabashi, 3,800-meter spark, received at McAllen, Texas.) In all cases, neither the character of the transmitter, the wave length employed, nor the local conditions or terrain have been effective in preventing the distortion of the electromagnetic wave used in signaling.

It may be said, however, that the measurements made when the transmitter employs an undamped wave of great length show distortion more frequently, and, in general, of a larger amount than those found when a transmitting system using damped waves and a shorter wave length is employed. It will be seen for instance, that XDA, Chapultepec, Mexico, low frequency spark using 5,700 meter wave length is but seldom distorted when observed at McAllen, Texas, while the measurements made upon the same station at Houlton, Maine, show more frequent and greater divergence from the normal; but still not to the same extent as has been observed in the case of NFF, New Brunswick, New Jersey, radio frequency alternator using

13,600-meter wave length received at the same station and over a much shorter distance.

It will be noted that the distortions found in the case of the long wave arcs during the daytime are of rare occurrence. It has not been possible to get any extended series of observations on short wave spark stations during the night. The ones taken, however, have shown but seldom any considerable distortion either by day or night. In the case of Tampico, Mexico, XAJ, spark system using 960-meter wave length shown in the figure, the distortion was very considerable, and was fully equal to that shown by the long wave arcs, taken, as was this, during the daylight hours. There does not seem to be any reason to believe that the short wave sparks may not, under unfavorable conditions, show distortion of the same magnitude as that given by other types of transmitters. In the case of Arlington, Virginia (NAA), using its 2,500-meter spark, there have been observed at Houlton, Maine, many distortions of large value and occasionally it was found that the loop receiver gave no directional indications. These distortions have been measured with great care, and the original data is given in one of the figures, showing a shift of 56 degrees 30 minutes in the angular direction of the signal wave within 13 minutes of lapsed time. There is also shown in the table a particular case where no directional reading could be obtained with a loop rotated about a vertical axis.

Other cases where no directional effect was observed are by no means rare, altho only a carefully kept watch would reveal all of them, since, as the illustrations show, the duration of this condition of no directional indication may persist for only a few minutes. During March and April, at Houlton, Maine, our records show this failure of the radiogoniometer on eleven different occasions. A good many other instances were recorded where there was a broadening of the minimum. The reason for this has been found in particular cases by other investigators to be caused by an apparent change in the wave front which may be corrected by a rotation of the loop about a horizontal as well as vertical axis. We do not know of any such examination when no directional effect at all is indicated by the radiogoniometer. All of the data obtained when no directional effect could be found show that two causes for the anomolous condition may exist.

First. Where the directional effect of the loop gradually disappears by a broadening of the minimum, but without any radical change in the directional indication: This would seem to show that at the receiver the resultant effect was just as tho the electromagnetic wave changed its direction of propagation from the horizontal to the vertical. Such a change in resultant direction would be shown by the establishment of disappearing points when the rotation is about a horizontal axis alone.

SECOND. At other times both before and after the disappearance of any minimum in the audibility, there has been found to be an extremely large and erratic change in the direction as indicated by the loop turned about the vertical axis. The data presented show particular illustrations of this condition. It would seem that the cause for the loss of directional effect must be of a radically different character from that in the first case. The explanation most probable is that at the receiver the electromagnetic wave is so modified as to produce a rotating field. When the two components of such a field have strictly quadrature relations, both with regard to time and with regard to direction, together with the same amplitude of oscillation, then the rotating field becomes purely circular and is the most simple case obtainable. It would be expected that this condition would persist only a very short time and would be preceded and followed by extremely erratic indications given by the direction of the loop. If this is the case then it would not be possible to obtain any directional effect by a rotation of the loop about the horizontal as well as the vertical axis. No experimental verification has been made in the case of long distance transmission, but such a condition has been observed by others in connection with radio experiments involving only a few hundred yards and employing very short waves.

The changes in audibility have also been obtained by an extended series of experiments, but the figures presented show only a few particular cases where it is desired to bring out the fact that large changes in directions are not necessarily accompanied by any considerable changes in the audibility of the signals and that even an entire loss of directional effect by the receiving loop may occur with no considerable change in the audibility of the signals.

The converse may also be true and it is shown that there may be very large changes in the audibility of the signals without any appreciable change in the direction.

In general, it may be said that the changes in audibility are of much more frequent occurrence and are proportionally much larger than the changes in direction. It seems, also, that the audibility changes follow a diurnal cycle as well as a seasonal cycle. Probably the underlying causes for the two changes are the same, but the directional changes only occur when the underlying cause produces a very special condition which does not have any considerable effect upon the energy absorption.

It might be expected, therefore, that most audibility variations would be produced by changes in the energy absorbed by the transmitting medium and more rarely by interference phenomena when the ionization causing absorption in general is so concentrated and segregated as to produce sharply defined conducting strata which would act as a reflector for the electromagnetic waves under conditions where the difference in path between a direct and reflected wave would be at most only a few half-wave lengths, and the angle would be small between the reflected and direct wave. The change in directional effect might be produced by the combination at the receiver of a direct and a reflected wave, but it would be necessary for the reflection to take place from a surface sufficiently removed in a horizontal direction from the receiver so that the reflected wave would have a very considerable horizontal component in order to cause the loop turning about a vertical axis to give an indication of a directional change.

In the second class of directional changes, it is assumed that there are two components, which have a quadrature relation both with respect to time and direction. This would require a reflected wave, as indicated above, from a surface at a considerable distance in a horizontal direction from the receiver, and at a distance which would make the reflected electromagnetic wave 90 degrees out of phase with the direct wave. If the distance were such that this phase difference was not 90 degrees, then the resultant wave at the receiver would show a direction depending upon the composition of the direct and reflected wave with respect to amplitude and direction. The phase combination of this reflected and direct wave might be almost anything, but would not affect the audibility of the signals as received by the loop, and so would not affect the general measurements upon which the direction indications depend.

We need now to examine with care the complex state existing in the atmosphere thru which the electromagnetic waves travel in order to determine the physical conditions upon which depend the observed phenomena of the direction changes and variations of audibility and strays.

The latest information that we have in regard to the conditions of the atmosphere shows that there are certain regions where there are discontinuities such as to make possible the formation of reflecting surfaces, assumed in our explanation of the observed phenomena. There are:

1. The Heaviside layer, assumed to be about 50 kilometers (30 miles) above the surface of the earth, and which has been discussed at considerable length by previous investigators.

There seems to be no doubt but that there would be a discontinuity and that it might very well happen that it would be much more pronounced during the night than during the daytime when the ionization produced by the sun's light might be expected to extend below the stratum of permanent ionization, diminishing rapidly as we proceed into the lower atmosphere. It might very well be that the greater transmitting distances observed during the night are obtained by the passage of the electromagnetic waves between the conducting surface of the earth and the conducting region of the Heaviside layer, which being a region of permanent ionization has a lower surface showing a sharp discontinuity, as soon as the ionization stops upon the disappearance of the sun and the ions present have rapidly recombined. There would, then, within this region below the Heaviside layer, be comparatively little free ionization and consequently but little absorption, allowing thus the transmission of radio signals over excessively long distances. During the daytime, the widely distributed ionization caused by the sunlight would result in large absorption of the energy, and this condition would be generally found to exist at all points. Irregularities in recombination at the disappearance of the sunlight might very well produce large regions in which the uniformity of the conducting layer is broken and where there persist for a time absorbing regions which would have considerable effect upon the audibility of the signals, and an erratic effect, as is usually observed in the case of the reception at night of radio signals.

The possibility of reflection from these large ionized masses or from the Heaviside layer itself in such a way as to produce the interference phenomenon which is indicated in the figures and which has long been known to the radio engineer as a "swinging" and "fading" of the radio signals at night, has been examined in detail by a number of engineers, and their conclusions will be found in the literature on the subject. It does not seem to us that it would be possible to have an interference produced by these reflections at the receiver, since the number of half-wave-lengths' difference between the two paths necessary in order to account for the observed phenomena would be so great as to make any sharply defined interference pattern out of the question, considering the nature of the reflecting surfaces and the characteristics of the transmitting medium. This would also be even more evident in the case of the changes of direction, since the distance would be very considerably greater on account of the necessity for having the horizontal component needed to explain the observed angular displacement of the direction of the wave. We must therefore look further for a region possessing the characteristics needed to produce the phenomena recorded above.

2. The diurnal vertical convection is limited to from five hundred to one thousand meters (1,500 to 3,000 feet) above the surface of the earth, depending upon the season of the year and the condition of the weather.

There would be a discontinuity which would appear at about the place where the measurements of penetrating radiation have shown that the effect of the radioactive constituents of the earth disappears. It does not seem to us that this would be a sufficiently well defined discontinuity to account for the reflecting strata needed to account for the observed phenomena.

3. There is, during the summer, a region which is defined by the cumulous clouds at four or five kilometers (2.5 to 3.1 miles) above the earth. We do not believe that this discontinuity is sufficient to account for the results.

4. There is, at approximately an elevation of seven kilometers (4.4 miles), a region known as the alto stratus layer, which is present in both summer and winter. This layer is not sufficiently well defined, and the causes for the discontinuity are not such as to lead us to believe that the observed phenomena can be accounted for by any reflections at this altitude.

5. There is always found a sharp dividing line between the troposphere and the stratosphere (isothermal layer) which occurs at an average altitude of ten kilometers (6.2 miles). This is dependent somewhat upon the season and also the storm areas. It is found that changes as large as three kilometers (1.9 miles) may exist between the cyclonic and anti-cyclonic regions. At this sharply marked boundary there is undoubtedly a layer of cosmic dust which may very well be strongly radioactive and so one of the principal causes of a stratum of permanent ionization. This is also the region where the cirrus clouds usually form, and which is consistently their upper

limit. It is the most sharply marked discontinuity of which we have any definite knowledge in the upper atmosphere.

It is probable that such a stratum is present during the night when there is little movement in the upper part of the troposphere and that this conducting layer forming the boundary between the troposphere and stratosphere, will act as an excellent reflector. The movement caused by a cyclonic disturbance will change, however, the inclination of this conducting layer. There may also be considerable regions in the part of the troposphere, bordering on its discontinuity which are strongly ionized, due to the upward drift of the negatively charged particles and water vapor which have been segregated in a sufficiently large quantity so as not to be immediately neutralized after the disappearance of sunlight. The cessation of the turmoil resulting from the convection currents caused by the sun's action will allow these ionized regions to assume a stratefied form which will act as excellent reflecting surfaces.

As soon as the sun reappears and convection currents are re-established, they both destroy the sharply defined conducting layer and cause a widely distributed ionization which absorbs much of the available energy. The result is a decrease in signal strength, but no considerable distortion in the direction of the transmission.

During the night the long distance transmission could take place between the earth and the Heaviside layer with the intermediate conducting stratum at approximately ten kilometers (6.2 miles) above the earth's surface playing but little part until some discontinuity caused it to act as a reflector. This conducting layer would also make possible abrupt local changes in energy due to absorption which would later, be equalized by a gradual distribution thru the conducting layer, which is of course not perfect, from the region lying between it and the Heaviside layer.

During the daytime, the greater ionization and its wide distribution below the Heaviside layer, as well as the constant occurrence of ionization below the stratosphere would both tend to absorb the energy of the electromagnetic wave, and so decrease the observed signal strength at the receiving station, while preventing the reflections necessary for the distortions observed during the night.

In the same way also the periodic change in audibility or "swinging" may be due to the movement of interposed conducting masses of air which absorb the energy of the wave. This would be particularly true when the periodic change occurs thru a considerable time. The interposition of a series of storm regions moving along the same direction would produce such changes in audibility. A test of this might be the determination of such variations with respect to a north-and-south line and an east-and-west line, since there is a decidedly greater storm movement along the east-and-west line.

There is consequently the condition necessary for reflection from the stratum which is approximately ten kilometers (6.2 miles) above the surface of the earth. The distance is such that it will permit the formation of interference patterns at the receiving station, since the number of half-waves would not be too large to make that possible. This might occur quite rapidly and be due to the interference of a direct and reflected wave when the reflecting surface is changing its position with a velocity often found in the movement of cloud masses. The entire disappearance of directional effect when due to the rotation of the electromagnetic field would also be easily possible when the reflection occurs from the ionized layer between the troposphere and stratosphere or from an ionized stratum below that boundary.

It would be expected that the changes in direction would occur, therefore, only when the discontinuity due to the stratification of the ionized gasses and particles became effective during the night hours, and that these changes in direction would be produced only by reflection and never to any appreciable extent by a refraction, so that changes of small amount might be expected to be of frequent occurrence, while changes of large amount requiring a reflection from a surface beyond the receiving station, so as to allow for the resultant direction caused by the combination of the reflected and direct waves to make a shift as great as the change of 56 degrees 30 minutes in thirteen minutes as shown, would be of much more infrequent occurrence than the audibility changes, continually observed, Also, the very special case where there is a complete disappearance of directional effect would not be produced except under very particular circumstances which might not be expected to occur except at rather infrequent intervals.

The results obtained have been quite disconcerting to those of us interested in making a quantative use of the directional characteristics of the loop, but it is believed that a knowledge of what may occur, altho but rarely, will be of material assistance in further work. It is possible also that our knowledge of the upper atmosphere may be greatly advanced by a systematic employment of radio frequency oscillations and the use of quantitative methods in radio measurements.

The observations from which the values plotted in the figures were obtained were taken by detachments of radio operators under the command of:

Captain James E. Ives, at Tucson, Arizona.

1st Lieutenant Alfred J. Main, at McAllen, Texas.

1st Lieutenant Arthur H. Boeder, at Houlton, Maine.

Many of the computations and the plotting of the figures was performed by 1st Lieutenant Lee Sutherlin.

The careful and interested co-operation of the officers and men engaged on this work, is most thoroly appreciated, and we take great pleasure in acknowledging our indebtedness to them.

SUMMARY: As a result of a large number of systematic measurements on direction of incoming signals, using loop reception, variations of considerable magnitude in the apparent direction and occurring within short periods of time are found. Their dependence on length and character of wave are studied.

There are certain possible theories of transmission proposed to account for the observed phenomena, notably reflection at the atmospheric "isothermal layer" (6.2 miles or 10 kilometers high), and the production of rotating fields at the receiving station.

DISCUSSION

Lester L. Israel: It appears that all of the loops experimented with it in the preliminary work were of the same form as that described in the paper which gave an asymmetrical polar curve for its directional characteristic. Accordingly, I would suggest a possible source of error in the experiments, which would lead to a different explanation of the variations in wave direction from any so far advanced.

It is well known to those advanced in radio compass work, that it is necessary to have a loop with an accurately symmetrical polar curve for correctly locating the direction of radio waves. Now the spiral type coil has a particularly pronounced asymmetrical curve as the preliminary experiments given in the paper show.

With this type of coil, the angle of maximum or minimum signal is dependent not only on the resonance tuning of the loop but also on the relative amounts of energy picked up by the loop acting as a loop and acting as an open antenna. It seems to me that this ratio would depend not only upon the amount of capacity unbalance, but also upon the dip in the direction of wave propagation or even the so-called slant of the wave front.

A wave coming from *the east* with vertical wave front and propagating horizontally might create a minimum signal with the axis of the loop 5° *south of east.*

A wave coming from *the east* with a 45° slant of wave front, as by reflection from overlying clouds or ionized layers, or by high resistance ground (dry weather), might create a minimum signal with the axis of the loop 25° *south of east.*

What I am particularly trying to suggest is that the lateral deviation indicated by such an unbalanced coil may really be a vertical deviation of the wave. Some such other cause must exist to harmonize with the consistent results obtained by other observers working with perfectly balanced loops.

Carl Kinsley: Many direction measurements were made using balancing capacities, but that method of measurement was not satisfactory for the investigation which was the subject of this paper.

Here it was desirable to use a method for taking the data which would correctly give the variations in direction which frequently took place with extreme rapidity. Only the horizontal component of the change actually occurring was noted in the recorded observations. The desirability of getting also the vertical component was pointed out in the paper.

Altho most of the work was done with long waves, a particular case is shown where the wave length was under one-thousand meters and the direction variations were ten degrees within an hour and the time of observation was during daylight when long waves usually gave correct directions. The short wave stations did not transmit during the night at the time these observations were made and so it was impossible to get the mass of continuous observations which were obtained in the case of long waves.

It can be said in conclusion that the method of measurement correctly showed the variations in directions and the large variations noted were almost always of short duration and came most frequently from midnight to four A. M. It is not likely that a single observer would be ready to note them. Only by continuous measurements over long periods of time was it possible to obtain the results reported in this paper.

NOTE ON RADIO FREQUENCY MEASUREMENTS*

By

CARL ENGLUND

(RESEARCH ENGINEER, WESTERN ELECTRIC COMPANY, NEW YORK)

Several articles on radio frequency measurements have lately appeared,[1] and, as considerable work of this kind has been carried out in the Western Electric Company's laboratories, it has seemed that a brief note,summarizing some of the results arrived at,might be of sufficient interest to warrant publication. This is the more so as bridge methods do not seem to be in general use elsewhere.

Alternating current bridges for radio frequencies have been in use in the Western Electric Engineering Laboratory for upwards of three years now, and a return to volt-ammeter methods of making measurements is unlikely. During this time, the laboratory measurements on coils, condensers, and circuits required for the radio apparatus manufactured by the Western Electric Company, have been made on these bridges. In addition the necessary measurements for the "multiplex telephone" development recently perfected have also been made in this manner. Naturally, for a great deal of the work, high accuracy has not been necessary nor desirable, but measurements to better than commercial accuracy have been frequent, and while the ideal bridge for commercial work, that is, one fitted with two terminals for connecting the unknown and appropriate dial handles for balancing to the requisite accuracy, is yet unborn (and, owing to the range in frequency and impedance to be covered is likely to remain unborn), this has not prevented the use of apparatus possessing a considerable convenience in the way of accuracy and rapidity with which measurements can be made. Lack of time and a feeling that such special apparatus, sometimes only used for the length of the job, might not have general interest, has so far prevented publication of the work done. Some general observations on an all-around bridge and the essentials for successful use of it may not be out of place

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* Received by the Editor, July 22, 1919.

[1] Dellinger, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, page 27, February, 1919; "Bureau of Standards Circular," 74.

here, however. Credit for the progress of bridge measurements is rather widely distributed among the past and present members of this laboratory, the radio frequency work being really an extension of the work by bridge methods begun some sixteen years ago when Dr. G. A. Campbell and his assistants introduced the shielded bridge, since so successfully used. At present the greatest trouble and chief source of error is in the standards used; it is possible to balance much more accurately than is justified by the standards available.

Any voltmeter-ammeter—or in other words deflection—method reduces to the reading of one scale. No greater accuracy is possible than that which the operator can attain by bisecting the smallest divisions with his eye. To expect more than 0.5 per cent accuracy for such an observation, near the top of the scale, would be optimistic, and a lesser degree of accuracy is usually attained. By balancing two deflecting instruments against one another in some manner, such as opposing two thermocouples, or winding two opposing windings on a dynamometer coil, we are enabled to use a more sensitive instrument and in effect have better than a single dial setting. But this is nothing more than a type of potentiometer work and with no standards of current or voltage available for alternating currents[2] as there are for direct currents (the Weston or Clark cell), it is logical to replace the unsteady balancing emf. by the drop across a known impedance and we then have the a.c. bridge. At once we reduce the balance to one of invariant (with time) circuit constants and free ourselves from the variations in current to which every radio frequency source is subject and, in addition, operate on very much smaller currents. Two milliamperes input is amply sufficient for any bridge measurement of low impedances (1,000 ohms, say). Another point is that no source of alternating current known at present is free from harmonics and the oscillating audion, the steadiest generator known, has them in plenty. This fact is not overlooked by a thermocouple, and tuning them out is difficult. The bridge balance can easily be made almost wholly immune to harmonics. Again, suppose a coil with ferric core is to be used in a receiving set. No thermocouple will operate on such small currents, and a measurement on larger currents gives incorrect results since the constants of the iron vary radically with the current used.

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[2] Alcutt, "Proc. Am. Inst. Elect. Engrs.," page 83, February 18, 1918, has proposed a method of establishing a secondary current standard applicable to alternating currents.

A bridge then becomes necessary. Only those who have had experience with such apparatus know what an advantage it is to be independent of a fluctuating source of radio frequency current.

Various types of bridges have been made and used in the laboratory here, and for special cases some have decided advantages. Thus, a capacity bridge for capacities only, and an inductance bridge for inductances only are very useful if expense can in some measure be disregarded. For all around work, the bridge later described has been most useful. Contrary to what might be supposed, radio frequency bridge measurements are not much more troublesome than measurements at audio frequencies. Up to something like 200,000 cycles, a bridge performs amazingly well. When 1,000,000 cycles is reached, trouble begins to be experienced, in measuring coils, due to the electromagnetic induction of the coil on detector circuits and the other bridge arms. This can be partly avoided by making the physical dimensions of the bridge small and spacing the generator, bridge, and detector. Shielding against electromagnetic induction has not proved very successful; but electrostatic shields, if made of material thicker than tin foil, leave nothing to be desired.

It is quite necessary, in general, that both generator and detector be either grounded at some point or completely enclosed in a grounded shield. With vacuum tube generators and detectors, if the negative end of the tube filaments (filaments usually in parallel) is grounded, no shielding becomes necessary. The generator inductance unit must be a toroidal coil, however; and since this will be a wood (in effect, air) core, the winding must be carefully done or an external magnetic field will be produced. This coil is easily subdivided into equal fractions, and with a switch may be thrown from "parallel aiding" to "series aiding" connections and thus greatly extend the frequency range. Two and three subdivisions have been used giving inductance ranges of 1 to ¼ and 1 to 1/9. Since the frequency should not vary as the bridge arms are altered at least two tubes are necessary for the generator. One tube oscillates freely, and the other tube, which is the power output tube, has the voltage thus produced impressed on its grid. With 150 volts plate battery and two "L" tubes, one should easily get an output of 10 milliamperes thru 1,000 ohms. By using more than two tubes, a generator may be made which has less pronounced harmonics but this is usually unnecessary. Figure 1 gives a generator circuit in use at present for the range 22,000 to 1,000,000 cycles.

If no terminal of the bridge may be grounded, two transformers will be necessary, for input and output, respectively. This is the case for a bridge used to make measurements on telephone or telegraph circuits where the effective impedance measured is center grounded. For work at commoner radio frequencies the bridge may be grounded at one corner, and only one transformer is necessary. The circuit diagram shown in Figure 2 gives such a bridge. For all around work, it is best not to include any standards in the bridge itself; these are best mounted in separate boxes and used as occasion requires. If any standards are to be included, a three dial resistance box is best. There is no objection to including standards in the bridge if only low impedance measurements are to be made, but they become objectionable for high impedance measurements. No advantage will be gained by having several sets of ratio arms, and resistance ratio arms are wholly preferable, both for the ease of construction and the fact that their impedance is independent of the frequency. They should be exactly alike and need not be more than 500 ohms in value. If they are wound with number

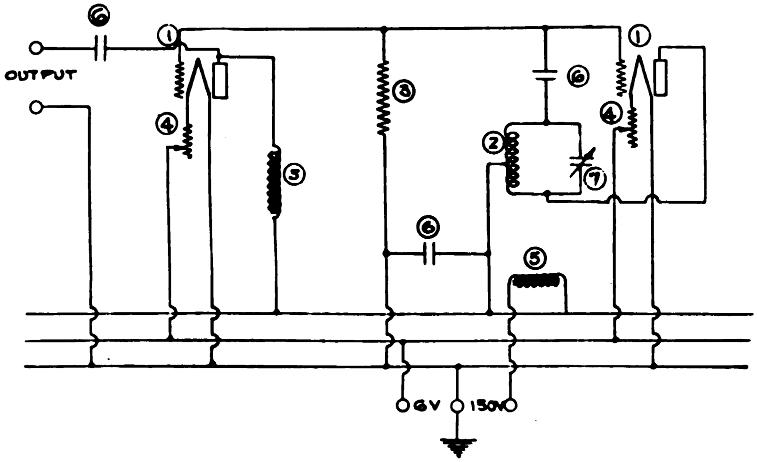

FIGURE 1—Radio Frequency Generator
① "E," "L," or "M"—Tubes
② Toroidal Inductance
③ 48,000 Ohms Grid Leak
④ 4-Ohm Rheostat
⑤ Radio Frequency Choke Coils
⑥ 1-Microfarad Paper Condenser
⑦ Tuning Condenser

38 B. and S. gauge*, or smaller, "Advance" wire in a single layer in either "reversed loop" or Curtis fashion, and then given a 0.75 inch (1.9 cm.) spacing from the surrounding shield, they will perform satisfactorily. The circuit of Figure 2 has double shielded ratio arms, the inner shield is connected to the "A" corner of the bridge and the outer is grounded. The capacity between the two shields thus bridges the generator and is without effect on the balance.

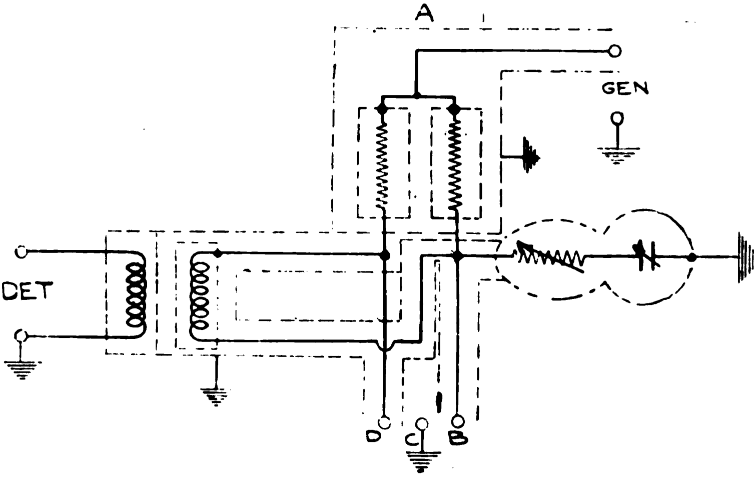

FIGURE 2—Utility Radio Frequency Bridge
(Shielding is dotted)

The output transformer is also double shielded, and preferably it contains a toroidal wood core and is mounted in an iron container. The distributed capacity between the primary winding and shield is without effect on the bridge balance since this balance implies that all parts of the primary winding and inner shields are at the same potential. The capacity between inner and outer shield is, in effect, in the "BC" arm of the bridge and a balancing condenser with either a variable low series, or a high shunt resistance, must be included in the "CD" arm to give the initial balance. For low impedance measurements this balancing member is unnecessary but it must be present when high impedance measurements are made. As a check on the bridge performance, we have usually included a switch so as to interchange the ratio arms and compare the two results. This is convenient but not absolutely necessary. It must be borne in mind that, at radio frequencies, the geometry of the layout must be considered and if a coil is being measured when so placed that its magnetic field generates considerable emf's. in the bridge network, trouble will result, The mutual inductance between primary and secondary in the transformer should be large in order to overwhelm small mutual impedances between bridge arms, detector, generator, and so on, and at the same time not exceed such values that the secondary may be fairly sharply resonant when tuned. For a range like 20,000 to a million cycles, a single transformer is somewhat unsatisfactory but normally preferable to any other arrangement since transformer coils (air core) with many taps are objectionable and two independent transformers, with switch gear, means too much apparatus. Except for such measurements as are, in effect, a severe test of the bridge, one transformer has been found good enough.

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* Diameter of number 38 wire = 0.004 inch = 0.01 cm.

The balance is always observed by using an auxiliary source, and producing a beat note in an ordinary telephone. A good detector will be obtained if one detector and two amplifier tubes are used. This, if designed to give maximum amplification, will give excessive and selective low frequency amplification (since the transformers become anti-resonant), which is undesirable. Less amplification over a greater frequency range is to be preferred. When the ear tires of one note another may then be used or the operator may select any frequency best suited to his ear. Only primary (or secondary) batteries have proved successful as sources of filament and plate current for the detector and amplifier tubes and to save dry battery renewals high "μ" tubes may be used instead of ordinary amplifier tubes and give the same amplification. Overloading is then easier, however. A noisy detector circuit usually means that either the plate or filament battery has run down or developed a bad connection. The greatest trouble likely to be encountered is that of false balances due to overloading the detector tube. Large harmonics or too large input of the auxiliary source will give false balances. It is not hard to guard against when this is recognized. The detector may be made to oscillate and thus act itself as an auxiliary source, but it is much better to separate the detector and oscillator functions and have a separate oscillator to give the beat note. The input from the auxiliary source is varied by moving the oscillator about on the table. Figure 3 shows a detector circuit.

The range from 5,000 to 15,000 cycles, when covered by using a beat note in the telephone receiver, gives trouble, as the harmonics of both generator and auxiliary oscillator are not far enough apart, and the operator easily gets "lost" in a jumble of notes.

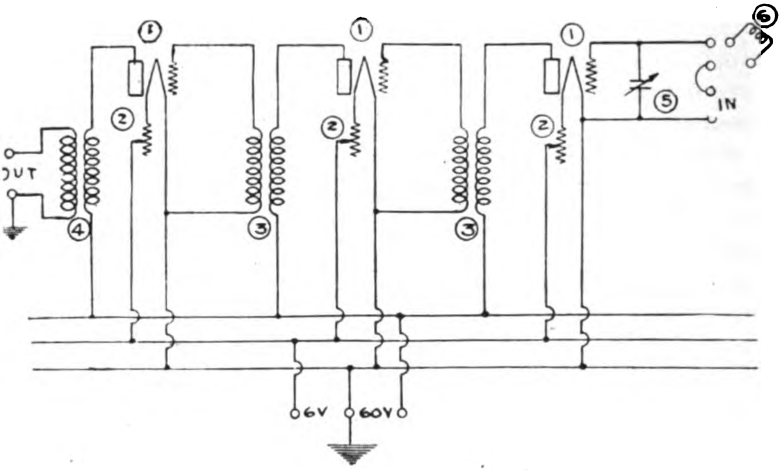

Figure 3—Detector Amplifier

① "D" or "V" Tubes
② 4-Ohm Rheostats
③ 61 A Retardation Coils Used as 1:1 Transformers
④ 201-A Transformer 400,000 to 20,000 Ohms (Use High Impedance Telephone)
⑤ Variable Condenser
⑥ Auxiliary Tuning Coils

Since most observers' ears are reasonably sensitive well over 10,000 cycles it has been found practicable to construct special telephone receivers and listen direct up to 10,000 cycles. Above this frequency, the beat method is used.

As stated, the limitations encountered are those of the stand-

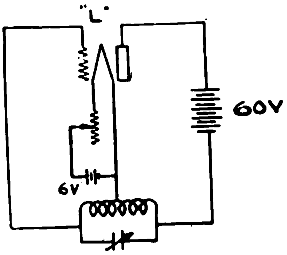

Figure 4—Heterodyne Oscillator

ards used. Settings giving a balance of one part in 10,000 are not difficult, but in the absence of standards mean nothing. Air condensers with negligible dielectric losses, calibrated at low frequencies; and carefully wound resistance units of number 38 or finer "Advance" wire*, and adjusted with direct current, have been so far used as the basis of all work. Coils are tuned with such a condenser and measured as a pure resistance. Frequency determinations are made by balancing a calculated coil, of negligible shunt capacity, with the standard condenser against a resistance in the bridge. Condenser measurements are usually made by tuning the condenser with a variable inductance, whose other properties are unimportant, and substituting the standard condenser and retuning. Direct balance of the unknown condenser against the standard is possible but the first method makes the balance a low impedance measurement which is usually an advantage. In determining the resistance of coils and condensers, a standard resistance may have some reactance without introducing any error. The so-called "shunt" capacity of coils is determined by successively balancing the coil with the standard condenser at two frequencies. The capacity then follows from the formula[3]

$$C_o = \frac{C_1 - \frac{\omega_2^2}{\omega_1^2} C_2}{\frac{\omega_2^2}{\omega_1^2} - 1}$$

Absolute frequency determinations should be made by bridging the gap from an audio frequency (determinable with a standard tuning fork or siren) and the desired radio frequency, by means of an oscillator of pronounced harmonics of high order.

It is expected that further results of this work will be published at a later time, which will throw additional light on the advantages of bridge methods as compared with deflection methods.

Research Laboratories of the American
Telephone and Telegraph Company and
Western Electric Company, Incorporated.

SUMMARY: After discussing the need for a radio frequency bridge of general utility, the bridge method is critically compared with indicating, null, and potentiometer methods.

Certain precautions in the shielding and coil construction of such bridges are considered; and the specific form of bridge used by the author is described in detail. Beat note with amplification and telephone indicator of balance point are used. Some applications of the bridge are then given.

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* Diameter of number 38 wire = 0.004 inch = 0.01 cm.

[3] Weinberger, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS volume 5, page 361, 1917.

NOTE ON THE INPUT IMPEDANCE OF VACUUM TUBES AT RADIO FREQUENCY*

By

JULIUS WEINBERGER

(RESEARCH ENGINEER, RADIO CORPORATION OF AMERICA, NEW YORK)

In the various papers which have been published on the behavior of vacuum tubes with three internal elements, one frequently finds the statement that the input impedance (the impedance of the grid-to-filament path) of the tube is infinite when the grid is maintained at a negative potential with respect to the negative end of the filament. That is, it is assumed that since the d.c. characteristic of grid potential against grid current shows that no current will flow between grid and filament when the grid is negative (in a high vacuum tube), the behavior with alternating, or radio frequency voltages, will be the same. While at audio frequencies this path may act as practically an infinite impedance, I have found that at radio frequencies conditions are quite different.

In tubes, as they are now built, there exists considerable capacitive coupling between the grid-to-filament path and the plate-to-filament path; this coupling is sufficiently strong at radio frequencies to cause the apparent input impedance of the tube to be a function of the character of its output circuit. For example, I have noted that if there is a resistance in the output circuit (as when the tube is used in a resistance coupled amplifier), the input impedance will depend on the value of the output resistance. If the output circuit consists of an inductance, the input impedance will act as a negative resistance, which will reduce the resistance of the input circuit (for example, the secondary circuit of a receiving set to which the tube may be connected as a detector) or even give rise to oscillations.

OUTPUT CIRCUIT CONTAINING RESISTANCE

In order to investigate the nature of these effects experimentally, the set-up shown in Figure 1 was made. A sustained wave oscillator operating at a wave length of 6,000 meters was

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*Received by the Editor, October 9, 1919.

located at a considerable distance from the set-up, and a pair of leads brought out to a single loop of wire, which was in turn, coupled to a second single loop. The coil L (about 9 millihenrys) and condenser C served as a receiving circuit to which the tube was connected. Loosely coupled to L was a circuit containing a coil, variable condenser, crystal detector, and portable galvanometer (the latter having a sensitivity of 0.1 microampere per scale division—one scale division equal to 2 mm.).

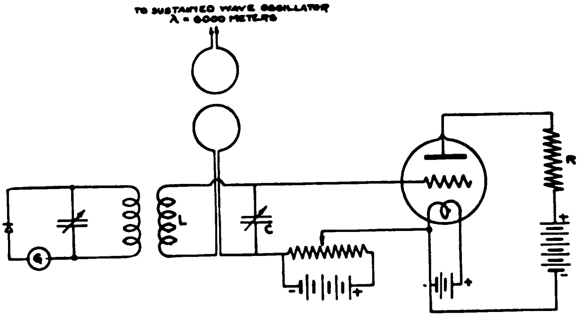

Figure 1—Experimental Set-Up

In series with the grid of the vacuum tube was placed a battery and potentiometer which served to maintain the grid at a negative potential with respect to the filament. The resistances R in the output circuit of the tube were of the deposited tungsten film type made by the General Electric Company.

In order to study the operation of the tube, the procedure was as follows: The tube was connected to LC, with various resistances in its output circuit, and the deflection of G noted for each case. A negative potential of 0.8 volt was maintained on the grid; the radio frequency voltage across C was always considerably less than this and no grid rectification was therefore possible.

After completion of the observations, the grid lead of the tube was disconnected from C, and the filament extinguished; then various resistances of the tungsten film type were shunted across C and the galvanometer deflections again noted. A curve of these is given in Figure 2. In all cases, the circuit $L\ C$ was kept tuned to resonance by adjusting C.

The equivalent resistance of the tube and its associated output circuit could therefore be read off from the curve of Figure 2 by finding the resistance corresponding to a given galvanometer deflection.

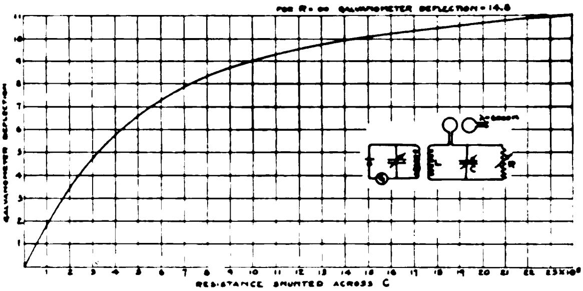

FIGURE 2

EXPERIMENTAL RESULTS

The results of these experiments are given in Table 1. It will be noted that as the output resistance is varied from zero to 5 megohms the apparent input resistance varies from infinity down to a minimum of 50,000 ohms and then up to 170,000 ohms. That this is sometimes equivalent to a considerable increase in effective resistance of the circuit *L C* is evident from the fact that a resistance of 2 megohms in the output circuit of the tube had the effect of apparently quadrupling the resistance of *L C*.

TABLE 1

Tube: Western Electric Company's Type D.
Plate Battery: 56 Volt; Filament Current: 1 ampere.
Grid Potential: 0.8 Volt Negative.
Wave Length: 6,000 Meters.
$L = 9.7$ millihenrys. $C = 0.001$ microfarad.

Resistance in Plate Circuit of Tube, ohms	Galvanometer Deflections Scale Divisions	Effective Input Resistance (Indicated by Galvanometer Deflection) ohms
0	14.8	Infinite
45,000	2.4	130,000
100,000	1.0	60,000
850,000	0.8	50,000
1,450,000	1.0	60,000
5,000,000	3.0	170,000

It is possible to explain the above results on the basis of the following theory:

For radio frequency, at least, a tube with resistance in its output circuit has an effective input resistance which is equal to the total plate circuit resistance divided by the voltage amplification given by the tube for the particular output resistance in question.

By "total plate circuit resistance" is meant the sum of the internal resistance of the tube and the output resistance; by "voltage amplification for a particular output resistance" is meant the ratio of the voltage obtainable across the output resistance to that impressed on the input terminals of the tube. The latter is measureable on a modified Miller bridge, as shown in Figure 3. The theory of the regular Miller bridge for measuring the amplification factor of a vacuum tube has been published.[1] This modification permits the measurement of the ratio of the voltage across the output resistance R to that impressed across the grid and filament. The operation of the apparatus of Figure 3 will be obvious from the diagram.

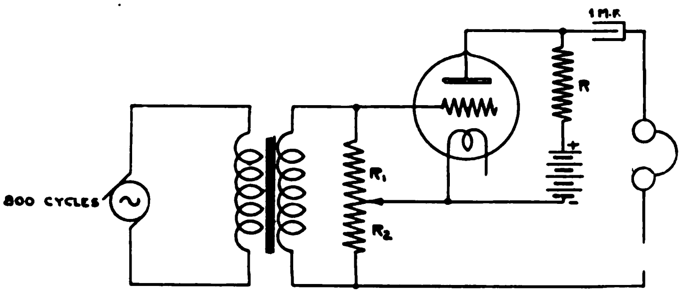

FIGURE 3—Modified Miller Bridge for Measurement of Voltage Amplification

In order to test this theory it was necessary to determine the total plate circuit resistance with various output resistances. This was done as follows: With a plate battery of 54 volts and the usual negative grid potential of 0.8 volt, a number of tungsten film resistances were successively inserted in the plate circuit. With each resistance the variation in plate current was observed on a sensitive d. c. galvanometer when the plate battery potential was increased and decreased slightly from the value of 54 volts (a variation of one or two volts was made).

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[1] John M. Miller, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 141, 1918.

The variation in battery potential divided by the corresponding changes in current gave the total plate circuit resistance. The values are tabulated in the second column of Table 2. In the third column are given the values of effective input resistance, corresponding to the output resistances of the first column, which have been taken from Table 1. In the fourth column are given the ratios of columns 2 and 3. These ratios correspond approximately to the voltage amplification values for the resistances of column 1, which are given in the last column.

TABLE 2

Resistance Inserted in Plate Circuit ohms	Total Plate Circuit Resistance (Measured) ohms	Effective Input Resistance (From Table 1) ohms	Ratio of Total Plate Circuit Resistance to Effective Input Resistance	Voltage Amplification for Resistance of Column 1 (Measured)
0	....	∞	0	0
45,000	220,000	130,000	1.7	1.8
100,000	325,000	60,000	5.4	5.2
850,000	1,050,000	50,000	21.0	23.5
1,450,000	1,860,000	60,000	31.0	28.9
5,000,000	5,450,000	170,000	31.2	31.0

GENERAL NOTES

The tube used for the above tests was a Western Electric Company's Type D. Similar behavior, at the wave length used, was observed with a Western Electric Company's Type J, and a General Electric Company's Type G tube.

Lack of time prevented the carrying out of extended experiments with inductive and capacitive output circuits and at other wave lengths. However, it was noted that an inductance in the output circuit of about 15 millihenrys caused a *diminution* of resistance of the circuit $L\ C$ to about one-fourth its value. That is, the effective input impedance of the tube was that of a negative resistance. On the other hand, shunting a resistance in the plate circuit by a condenser of 0.001 microfarad, caused the effective input impedance (which previously had been a low resistance) apparently to become an infinite resistance. That is, a capacitive output circuit produces neither a positive nor negative resistance effect at the input terminals.

Research Department of the
Radio Corporation of America,
November 24, 1918.

SUMMARY: It is shown experimentally that the input impedance of a three-element vacuum tube at radio frequency is not infinite, as heretofore supposed, but that because of the capacitive coupling between input and output circuits within the tube itself, the input impedance greatly depends on the character of the output circuit. Experiments with resistance in the output circuit are given and a theory is advanced to explain the results observed. The general effect of inductive and capacitive output circuits is mentioned.

DISCUSSION

Lewis M. Hull (by letter): Mr. Weinberger's paper shows his appreciation of the importance of a matter which will, in the future, surely figure largely in the intelligent design of all electron tube receiving apparatus. However, I wish to comment on his method of measuring the input impedance. It seems that the effect upon the current thru L (Figure 1), which is what is indicated indirectly by the galvanometer deflection, of connecting a pure resistance across the condenser C, is intended to replace entirely the effect of the input impedance of the tube that in a similar connection, or at least the resistance component of that impedance. Apparently it is assumed that the input impedance is a pure resistance; the statement that "the circuit LC was kept tuned to resonance" does not refute this, since the addition of a parallel resistance would detune the circuit. Attention is therefore called to two points in this connection. (In this discussion C_1 = capacity between grid and filament, isolated; C_2, grid and plate; and C_3, filament and plate.)

(1) This method does not indicate the total impedance of the tube, as may be deduced from the observed result: "with the external plate circuit resistance equal to zero the effective input resistance is infinite." The input impedance under these conditions is really C_1+C_2. Taking average values for a "D" tube and socket ($C_1 = 12\ \mu\mu$f.; $C_2 = 18\ \mu\mu$f.), this makes the impedance at 6,000 meters, 106,000 ohms, which is less than two of the resistance values given in Table 1. This impedance, being wholly reactive, and representing a detuning of the capacity C by only 3 per cent., would, of course, give a deflection of the galvanometer closely equal to the limiting resonance value, attained when R is infinite.

(2) Likewise the galvanometer deflection obtained with the tube connected across C, translated in terms of the R in Figure 2, does not indicate the resistance component of the input impedance. It has been shown ("Dependance of the Input Impedance of a Three-electrode Vacuum Tube Upon the Load in the Plate Circuit," Scientific Paper of the Bureau of Standards, Number 351) that the input impedance of a tube having a pure resistance load in the plate circuit can be represented by a resistance in series with a capacity. Over a wide range of radio frequencies, both the resistance and capacitive components of the impedance are important. At the wave lengths employed by Mr. Weinberger, the input capacity is given by

$$C_g = C_1 + C_2 \left[1 + \frac{\mu R_p}{R_o + R_p} \right]$$

which becomes C_1+C_2 when R_p is short-circuited. The fraction in the brackets is, of course, the over-all voltage amplification. Thus it would seem that the phenomena in the measuring circuit are more complicated than is implied by Mr. Weinberger's discussion. If a certain reading be obtained at resonance in the *LC* circuit, for a given value of the resistance *R*, it must be apparent that for a series combination of resistance and capacity giving exactly the same deflection, *R* does not necessarily represent either the effective resistance or the total impedance, particularly in view of the fact that the capacity-reactive component of this impedance is usually larger than the resistance component.

Julius Weinberger (communicated): Mr. Hull's discussion brings out a number of points which are quite correct. Dr. Miller has, no doubt, covered the theory of operation of a vacuum tube in the admirable paper cited by Mr. Hull, in a much more thoro manner than the foregoing note of mine. However, the data given in this note was taken about a year before Dr. Miller's paper appeared and was written up at that time, but owing to other questions involved, publication could not be permitted until recently.

It is possible that the method employed in these experiments will prove useful to investigators who are limited in the amount of apparatus at their disposal. Dr. Miller's method, using a capacity bridge at 1,000 cycles, is possibly more accurate in its determination of the various capacities which enter into this phenomenon; but even Dr. Miller states in his paper that the dielectric losses in the tube capacities exert an important effect on the input impedance, and this in itself would indicate that a radio frequency measurement is preferable (since it cannot be assumed that the dielectric losses at radio frequencies can be computed from audio frequency observations). Moreover, the determination of the capacity component of the input impedance is quite possible with the radio frequency method by computation from the observed amount of detuning on the variable condenser. When a tube is connected to the secondary circuit of a receiving set, the resistance component is all that is important since the capacity component (which may just as correctly be assumed as being in parallel with the resistance, as in series) is then compensated for by the detuning of the variable condenser. The radio frequency method here shown then indicates directly what happens in the receiving circuit.

PROCEEDINGS OF
The Institute of Radio Engineers

Volume 8 OCTOBER, 1920 Number 5

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PUBLISHED BY
THE INSTITUTE OF RADIO ENGINEERS, INC.
THE COLLEGE OF THE CITY OF NEW YORK

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ALFRED N. GOLDSMITH, Ph.D.

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AN ELECTRICAL SIGNALING METHOD FOR GUIDING AERIAL AND MARINE CRAFT*

BY

ROBERT H. MARRIOTT

(EXPERT RADIO AID, BREMERTON NAVY YARD, SEATTLE, WASHINGTON)

It is evident that vessels have not had means for sufficiently accurate guiding of them in narrow channels, dangerous channels, and in fog. Largely for this reason, many vessels have been lost or delayed. For example: one hundred and sixteen vessels have been lost in Alaskan waters north of Ketchikan. Many vessels are held up for considerable periods of time at harbor entrances, for example, at New York harbor entrance, where the financial cost for delay may amount to as much as $500 per hour for one vessel. It is believed that recurrence of such losses may be prevented by the following method.

Install a signal-carrying conductor along the sides of the channel or under the channel and pick up the signals on the vessel. For example, when conductors are on each side of the channel, guide the vessel so that the signals from the starboard conductor are slightly louder than those from the port conductor. Also, when two conductors are under the channel, one for inbound and one for outbound, keep the inbound vessel in the course that gives loudest sound indicating the inbound guide. The under-channel guide would probably be more practical.

The ship receiving device may consist of a coil of wire on an iron core attached to the bottom of the ship with the core athwart-ship for receiving from a submarine signal wire or vertically to the side of the ship for receiving from a wire parallel to the channel. The winding of the coil may be connected to a pair of telephone receivers worn by the helmsman or navigating officer, or one pair of telephones for each of the officers.

Where conditions will permit, the signal may be the spoken word "In" for the inbound channel and "Out" for the outbound channel.

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* Received by the Editor, July 14, 1919. Presented before the Seattle Section of the Institute, November 5, 1919.

If desirable and conditions permit, different frequencies of alternating or pulsating current may be impressed in the guide conductors,—one frequency for inbound traffic and one for outbound. These signals may be picked up by both telephones and a signal-strength indicating device (as, for example, a sensitive ammeter or galvanometer in a circuit resonant to the frequency of the current impressed on the guide conductor).

With two or more such indicating devices, the relative strength of the fields of the inbound and outbound signals and the relative strength of fore-and-aft and athwart-ship fields may be ascertained. Also by controlling the strength of the signal in accordance with the distance from a given point, for example, by grounding a portion of the current thru suitable resistances at points at fixed distances from the source of supply, the strength of the signal would be dependent on the distance from that source. Such indications of strength of signal relative to the source of supply (for example, the dock) and relative to the inbound and outbound signal conductors and relative to the angle the ship makes with the signal conductor, offer means for a relatively high degree of accuracy in steering the vessel so as to keep clear from danger.

A further complexity of circuits may be provided which will afford sufficient accuracy in certain places, for example, the guide may consist of several conductors in one cable. The conductors would be of several suitable lengths and suitable speech currents would be impressed on them. For example, one conductor would carry a speech current which says periodically "One mile in," while the next in length would say "Two miles in" while the outbound signal wire in that vicinity might say "Nine miles out."

The ship's receiving circuits may be arranged fore and aft and athwart-ship in such manner as to indicate whether the ship is approaching the signal conductor, or leaving it, or is parallel to it, or at some angle to it.

Structural conditions of ships, channel conditions, and financial and political conditions have their bearings on what structural arrangements would be best to carry out the method.

For example, in New York harbor it would probably be entirely practicable to provide all possible refinements and strong accurate signals which could be followed by vessels equipped with the more simple inexpensive apparatus as well as those which might be equipped with more expensive and more highly accurate apparatus.

In the long passages in Alaska, it may be practicable and sufficient for the present to follow the Army cable lines. This may be made possible by picking up the telegraph signals and guide signals which might be supplied by the Army in intervals between telegraph signals. To do this is more difficult than would be the case if the cable were made for directing purposes, because the Army cable currents are weak and the cable is covered with a sheath containing iron. Also the resistance, inductance, and capacity of the long cables limit the quantity and kind of signals that may be impressed on them.

However, there are several methods which can be used as may be necessary to pick up the weak cable signals altho they are in an iron sheath.

For example, the pick-up circuit (which may be relatively small) may be lowered from the ship to near the bottom and maintained in the same relative position relative to the ship; also prescribed frequencies may be used on the cable and tuned receiver circuits and very weak currents may be amplified.

The receiving or pick-up circuit may include a coil virtually wound around more or less of the hull; one coil, for example, wound in a fore-and-aft plane and another in athwart-ship plane, or a coil or coils inside or above the hull depending on the practicability of these various arrangements.

Increasing the efficiency of the pick-up coil, increasing the current in the signal wire, and increasing the amplification all serve to increase the possible distance.

Only one signal conductor can serve, if necessary. For example, if vessels are to pass to starboard in a channel, two receiving circuits (port and starboard) may be used as indicated above, and the vessels steered so that the sound is always louder on the port side.

Portions of the guide conductor may be enclosed in iron for reducing the signal strength at desired points or for desired sections. Such an arrangement might prove desirable for certain of the conductors where several conductors were used; for example, one conductor might have the words "Port ninety" for a sharp turn and be in iron beyond the turn.

A first-rate, well-protected signal cable containing more than one conductor would probably cost a-dollar-a-foot, laid. Less expensive cables could be laid, however, depending on the depth of the water, the channel currents, and whether the bottom was soft or abrasive.

It may be desirable to arrange the incoming and outgoing guide conductors as a circuit instead of having a ground return for each, or it may be desirable to have a common metallic return conductor for both. The return conductor might be bare wire. Some conditions might permit the use of a bare wire conductor on shore on insulators with a return bare conductor circuit under water. The bare conductor might be of such low resistance that it would be the principal current path and so serve as a guide cheaper than where under-water insulation was used.

The signal current in the conductor along the channel may be like that in the antenna of a radio transmitter, and the radio receiving outfit on shipboard may be used to receive it, using a regular antenna or a coil. Also the coil might be rotatable on its axis so it could be turned to pick up the signal and then turn the boat so it would have the proper fore-and-aft position relative to the coil and therefore be headed right in the channel.

At this date it is hard to say what is the most practical arrangement to use to suit conditions. It may be best to go along the lines of making the best direction control apparatus irrespective of the radio apparatus and leave the radio apparatus free to handle other things, or it may be best to impress common radio frequencies on the signal conductor or load the antenna for lower frequency so that all vessels equipped with radio could tell when they were closest to the conductor by the strength of the receiving signal. The Alaska cable is probably not suitable for using radio frequencies.

However, the use of the common radio frequencies in the guide conductor might interfere with some stations which did not want to receive direction signals but did want to receive messages from distant stations, and it is easier to build under-water conductors for longer wave lengths. This indicates the use of the lower radio frequencies.

Fifteen thousand meters is about as long a wave length as any kind of a radio station receives at present. Therefore, we might go to a slightly longer wave length, that a ship's antenna could be loaded up to, and which a closed coil could be built for, and would be rotatable for getting the direction not only of this conductor but of radio stations. All such signals could be received on detectors, interrupted properly if sustained wave.

Suppose we take 20,000 meters, which corresponds to a frequency of 15,000 cycles. Fffteen thousand cycles is audible, but our ears do not pick it up well. Training might, however, make 15,000 cycles a practical frequency to hear.

Fifteen thousand meters probably could best be impressed on a submersed insulated wire about 3,000 meters or two miles long if the wire were grounded at the one end—probably the end near the transmitter generator. That frequency probably would do well in even longer wires or in a succession of wires one mile long, inductively or otherwise coupled at the ends; in which case the ship in passing over the points of coupling would note differences every mile which would be used as a measure of distance and further contribute relative to location.

All things considered, possibly the most practical form for common service would include a standard frequency for all ocean-going vessels and for all ports, of say 750 cycles, and enough current in the signal circuit so that the receiving circuit would not necessarily include anything but a rotatable coil on top of the pilot house, a fixed condenser, and a pair of telephone receivers with an indicating dial and coil-turning handle in the pilot-house.

Such a frequency is applicable to quite long wires, and indeed quite long wires might be required in some places. Furthermore, such a sinusoidal frequency and current can be fairly easily obtained from an alternator and maintained with very small percentage variation. Then, to such a receiving coil and condenser and receiver can be fairly easily maintained in comparatively constant adjustment.

Sinusoidal current and the frequency of 750 or 1,500 would serve to cut out some interference. Those frequencies are different from the commercial 60-cycle current, and the radio 500- and 1,000-cycle frequencies also seem to be sufficiently different from common motor and dynamo commutator and field frequencies. The average ear is probably most sensitive to about 750 cycles, but quite sensitive to 1,500 cycles.

Special vessels (for instance, ferries) which do not use regular channels probably should be provided with additional special frequencies. On the Alaskan cable route, special very low frequencies (maybe twenty cycles or less) would be more practical. It might be found that the longest Alaska cable run has a minimum impedence for a frequency of four cycles or less.

For airships and airplanes, the guide conductors probably could be established on existing pole lines and suitable frequencies used with certain dot and dash signals as for water craft; or, by use of amplifiers, speech signals might be impressed on the guide wires, using modulated direct current or suitable alternating current to which the airship guide receivers or other receivers were tuned.

There seems to be sufficient number of applicable forms, modifications, and combinations to provide a standard arrangement for each group wherein the service and service conditions have a sufficient number of common characteristics.

With the above descriptions and discussion, a brief description of one experiment is probably sufficient.

A 4,000 circular mil (0.02 sq. cm.) rubber-covered wire about 700 feet (214 m.) long was dropped over into the water along the side of a dock. One end of the wire was bare and in the water, the other end was brought up on the dock and connected to one side of the "break" on a buzzer. The other side of the "break" was connected to the water. Five dry cells were used to operate the buzzer. The wire along the dock under water and the ground return showed a bridge resistance of about 5,000 ohms. The predominant frequency produced by the buzzer corresponded to about 1,200 cycles. The rubber-covered wire in the water was intended to correspond to a short guide conductor in a channel, the center of the channel being along the dock.

The receiving device consisted of a wooden frame four feet (1.2 m.) square wound with two coils of number 28 double silk-covered wire* (L and L' in Figure 1). The coils were side by side separated by about 4 inches (10 cm.) each coil having a winding space cross section of about 0.5 inch (1.2 cm.) diameter. The coils were connected in series and had a resistance of about 1,000 ohms and inductance with telephones of about 10 henrys. There were about 1,000 turns. A variable air condenser of 0.005 microfarad maximum capacity (C) and a pair of 2,200-ohm telephones were connected in series with the total coil.

Wearing the telephones and carrying the condenser and coil (as shown in Figure 1) along the length of the dock with the coil vertical to the dock and its lower 4-foot (1.2 m.) side parallel to the signal conductor, the buzzer signals were sufficiently loud to be heard above the noise from riveters near the dock, the wind and induction from commutated circuits, and so on.

A tug, with steel hull and deck houses, was lying by the dock, its side being directly above the signal wire. The receiving apparatus was taken to the top of the pilot house, and the same results obtained.

The signals fell off as the coil was revolved away from parallelism to the guide conductor. No buzzer signals were heard when the plane of the vertical coil was at right angles to the conductor. Tipping the coil from vertical indicated on which side the conductor was located.

The signals could be tuned in or out by adjusting the condenser.

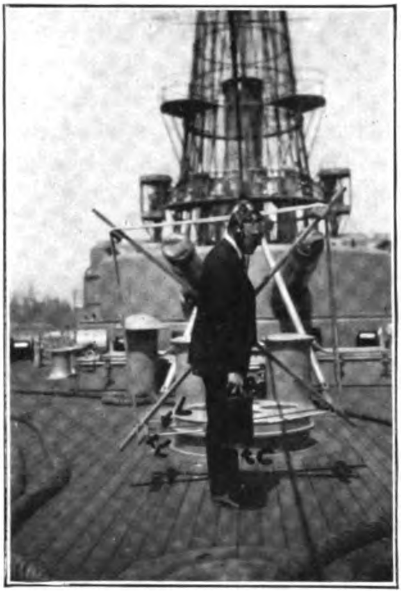

FIGURE 1

The above experiment can be easily duplicated or approximated, and indicates that a tug at least can be accurately guided by this method in a channel, which is deep enough for a tug. And the experiment, together with the previous descriptions, probably will indicate to those experienced in radio, telephone, or alternating current electrical engineering, that the method with suitable variations, somewhat as outlined, can be applied to various service conditions over comparatively great distances.

The cost of a suitable cable cannot be stated in general terms. Local conditions and the length of the channel may be such that the cable should be inexpensive both for first cost and maintenance. A five mile (8 km.) cable we have used for steering a vessel along the lines described, cost about $800 per mile; while on the other hand more profitable service might result by installing cable that would cost $10,000 per mile. In New York Harbor, the most suitable cable arrangement might cost a great deal and pay for itself in the first foggy month.

There are a wide variety of arrangements which can be made to suit the peculiarities of harbors, vessels, traffic, and so on. There seems to be a practicable solution for each peculiarity. For example, an unequipped foreign vessel may be served by the harbor pilot using receiving apparatus that he can carry with him, providing the signal cable current is made sufficient.

Sharp turns can not only be taken care of by the method previously mentioned, but also by shifting the cable from the true channel so the magnetic trail as indicated by the position of the receiver, will indicate the true channel.

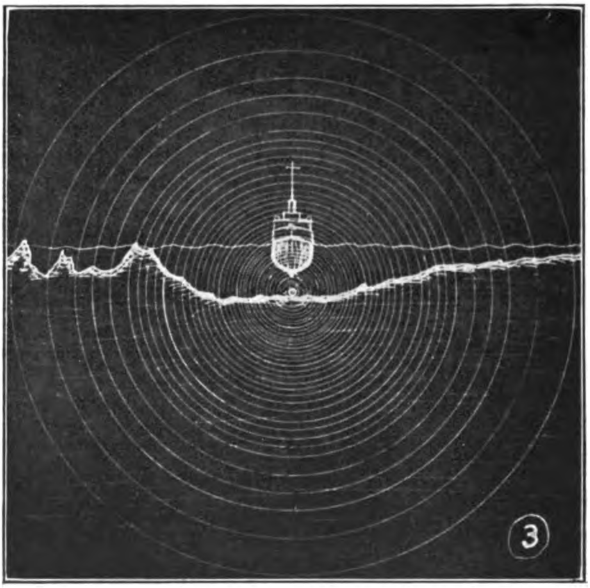

Figure 2

With audio frequency apparatus, any interferences are due to current variations on the ship and, being of nearby origin, it is possible to neutralize such local disturbances by inserting a small coil in series with the main coil and bringing the small coil near and properly placed relative to the local source of disturbance so the disturbance picked up by the guide coil is neutralized without neutralizing the signal. By proper allowance for phase relations this may also be done by inductively coupling the neutralizing system to the guide coil. Sufficient current in the cable and selected frequency also serve to minimize disturbances.

It is also more or less apparent that if sufficient current is used in the cable, the vessel might be automatically guided. One or two schemes have been designed for this during the consideration of providing a suitable automatic device which will give the pilot a visual indication of where the ship is relative to the guide.

It will also be apparent that along these same lines a cable could be laid along a defense line and a submarine could ply back and forth under water along such a defense line and, not only be guided by the cable, but could handle messages to and from the cable. Also that automatic torpedoes or mines could be moved along such a cable by a control station. The control station could, for example, learn from airplanes where the enemy was going to cross the cable.

The receiving coil, coil supports, and the location of the coil may be such that the same coil frame can carry a smaller inductance coil for use as a radio compass. With this combination the coil arrangement can be used on the high seas to find the location of radio stations on ship or shore and thereby steer for the entrance of the channel. Having reached the entrance of the channel, the simple throwing of a switch is sufficient to change the equipment to pick up and follow the cable guide signals.

The first complete practical installation was made on the S. S. *Tourist* of the Kennedy Line, Navy Yard Route, using the United States Army submarine cable from Bainbridge Island to Pier 8 in Puget Sound Navy Yard.

Captain Barrington of the *Tourist* has repeatedly guided his vessel by sound over this cable; and on one or more occasions the results indicated that it was easier to guide the vessel over the cable by this sound method than by the buoys, even when the buoys could be seen at a considerable distance. This may be because the drift due to tide currents is more noticeable by the channel cable electrical signal method than when trying to keep in line with buoys.

The cable was loaned by courtesy of Colonel Lenoir of the United States Army Signal Corps and Lieutenant-Commander Luckel and Gunner Thomas of the 13th Naval District Communication Service. The use of Kennedy Line Vessels was obtained by courtesy of Captain Mitchell, Manager. The writer was assisted by Mr. Schoenborn, Mr. Rowe, Mr. Palmer, Dr. Lester, and other members of the 13th Naval District Radio Engineering Organization.

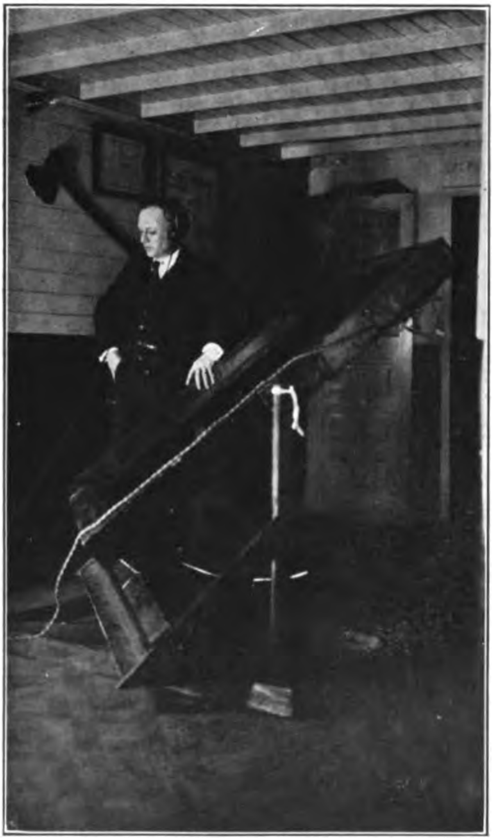

FIGURE 3

The cable is an ordinary submarine telegraph cable. The conductor is covered by insulation and a sheath of iron wires.

One end of the cable conductor was grounded at Fort Ward and the other was connected thru a key, ammeter, resistance and a 500-cycle generator to ground at Puget Sound Navy Yard Radio Laboratory. A 500-cycle generator, resistance, ammeter and key was also placed on Pier 8 for use at that point when more convenient. The morse signal "M A" was used as the guide signal.

From the standpoint of providing a wide margin of safety, the potential applied to the cable was limited at 40 volts. With this potential, the maximum cable current obtainable was seven-tenths of an ampere.

The receiving coil consisted of four coils of number 28* double silk covered wire connected in series, and in series with variable air condenser having a maximum capacity of 0.005 microfarad and the primary of a vacuum tube amplifier circuit. The system showed minimum impedance for 500 cycles when the capacity was about 0.004 microfarad. There are about 2,000 turns of wire in this total receiving coil and the coil is about four feet (1.2 meter) square.

Three steps of amplification were used to pick up the signal from the 28 watts or less in the cable which is about five miles (8 km.) long.

For convenience, the receiving coil and amplifiers were placed in the smoking room, under the pilot house and a pair of receiving telephones were connected for use of the pilot in the pilot house. Other telephone receivers were connected in series for use in the smoking room. With this arrangement it was necessary for the Captain to keep the vessel within about three hundred to six hundred feet (100 to 200 meters) of the cable to hear the signal.

As the pilot could not see the receiving coil, the coil was usually left vertical and fore and aft and the Captain swung the ship as necessary to hold maximum signal.

The receiving coil was turned and tipped for getting the location of the ship relative to the cable, as shown in the photographs. At times this was done when the vessel was being operated and the location signalled to the pilot. The coil is supported on pivots so it can be turned around a vertical axis or tipped around a horizontal axis. The loudest sound is heard when the coil is parallel to the cable and entirely on the same plane. The least or no sound is obtained when the coil is at right angles to that position.

In Figure 2 are shown the magnetic lines of force of the current in the cable. From their directions, the mode of operation of the system can be readily deduced. The actual receiving coil used on the *Tourist* swung out of its normal positions relative to its vertical and horizontal axes, is illustrated in Figure 3.

The harbor guide circuits should be so constructed, maintained, and operated, as to guarantee uniformity and reliability.

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* Diameter of number 28 wire = 0.0126 inch = 0.032 cm.

Therefore, their construction, maintenance, and operation should probably be defined by Federal law as soon as sufficient experimental and development service has been rendered to indicate just what complete arrangement or arrangements will render the maximum service. Too early definition by law might serve to handicap development.

Development by different private companies and by government departments would probably result in a number of different forms and installations which would later have to be standardized into certain lines of uniformity to render the best service.

Probably the method for development and application which would provide the most efficient service in the shortest period would be for some department of the government to provide a special organization for this work, under appropriations from Congress.

The business of this organization should be simultaneously to study the service requirements and carry on experiments and development with apparatus of government design, and such as might be purchased from private interests that might give their attention to this work. The development experiments should then be followed by the establishment of permanent standard guide circuits and by laws and international arrangements as necessary; probably leaving the supplying of suitable receiving apparatus to private concerns which might rent or sell them to the steamship companies, operating under various flags.

Radio compass stations are already being established. These can be used to guide vessels to the harbor entrances, where the guide signals could be picked up and followed thru narrow or crooked channels and past outbound vessels without seeing channel markings or the vessels and without having previously navigated that channel.

Since the writing of the above, several articles have appeared somewhat along the same lines. In those articles the matter of invention has recsived some attention. To make the above paper more nearly complete some remarks relative to patents probably should be added.

My files, in which the more basic inventions along this line are noted, have been in storage since 1914; however, it is my recollection that a patent was issued to Thomas A. Edison in 1878, which quite broadly covers the principle of intercommunication between a moving object (for example, a steam train) and a conductor along the line of travel.

Next in importance, probably, is the patent which was issued to Robert E. Owens in 1903 and of which the following is a brief extract:

> "No. 736,432, Apparatus for Ascertaining Position Relative to a Pre-arranged Guiding System. Robert E. Owens, Montreal, Canada. Filed March 11, 1902. Serial No. 97,799, issued August 18, 1903.
>
> "Claim 1. In apparatus for ascertaining position relative to a pre-arranged guiding system, a submerged electrical conductor located along the course to be followed by a vessel, a shore-circuit connected with and supplying the submerged conductor with a current adapted to create a magnetic field within which the vessel moves, and a double-coil indicating device carried by the vessel, adapted to be affected by said magnetic field, whereby the vessel's position relative to the submerged conductor may be ascertained."

SUMMARY: Inability to guide vessels accurately results in large losses of life and property. An arrangement by which vessels may be guided accurately will have a value corresponding to what it saves.

A signal-carrying conductor along the course of a vessel, and means capable of indicating the proximity and direction of the signal conductor are described and discussed, including various forms and combinations. Ocean-going vessels entering harbors are chiefly considered, the same principle being applicable to other craft and channels, air ships, and so on.

Experimental tests of the system are described in detail, with numerical data; and suggestions are made as to how the system may be further developed and applied.

STATIC ELIMINATION BY DIRECTIONAL RECEPTION*

By

GREENLEAF W. PICKARD

(CONSULTING ENGINEER, WIRELESS SPECIALTY APPARATUS COMPANY, BOSTON, MASSACHUSETTS)

In the early days of radio communication, those disturbances which have been variously called "static," "atmospherics," "X's" and "strays" were either attributed to distant lightning, or more vaguely and correctly to "atmospheric electricity." While it is still true that our knowledge of static causes is far from satisfactory, we have at least progressed to the point where the distant thunder-shower has ceased to be considered an important source of static, and are now able to concentrate our attention on the higher levels of the atmosphere. Such disturbances as may exist at these levels are unquestionably electrical, so far as they concern our receiving circuits, and are due to a supply of electricity which in some way is generated in or supplied to the atmosphere. The normal electrification of our atmosphere, observed as a rather marked potential gradient in the lower and more accessible layers, accompanied by downward electrical currents and attended from time to time by high level auroral displays and magnetic storms, is undoubtedly due to ultra-terrestrial sources. According to Arrhenius[1] the sun is the ultimate cause of electrical phenomena in our atmosphere, and altho his original theory has suffered certain modifications and additions in the past fifteen years, it will well serve as an introduction to this paper.

The sun being an intensely heated body must in consequence have emitted electrons in such abundance as to have acquired a very large positive charge. This charge, estimated by Arrhenius as three billion volts, forms a powerful center of attraction for stray electrons and negative ions, which are gathered in even from space far beyond the confines of the solar system. An electron or a negative ion, moving in toward the sun under this

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*Received by the Editor, October 14, 1919. Presented before THE INSTITUTE OF RADIO ENGINEERS, New York, November 5, 1919.

force, will eventually reach the outer and cooler solar atmosphere, where it will form a nucleus for condensation. A droplet, perhaps consisting of iron or calcium, will grow around this nucleus until it attains a diameter of the order of a wave-length of visible light. At this point in its growth, the pressure of light upon its surface considerably exceeds the solar gravitational pull upon its mass, and the droplet moves away from the sun. Such of these minute particles as may reach the vicinity of the earth fall into our upper atmosphere, and as they carry with them their original electrical charges, they constitute an important source of electricity. There are, however, two other sources, also solar, but differing materially in their mechanism. One of these is the ionization produced by the shorter wave-lengths of the sun's light, as it impinges upon the gas atoms in the upper strata. The other consists of a direct emission of either electrons (Beta rays) or, according to Chapman[2], charged helium atoms (Alpha rays). This direct emission apparently does not take place all over the sun's surface, but jets out in streams, presumably most vigorously from sunspot areas, in consequence of the great thermal gradients and magnetic forces existing at such spots. When such a stream, which may be ten or more degrees wide, sweep across the earth, it makes itself evident as magnetic storms and intense aurora.

The aurora, or Northern Lights, give us a certain illumination as to the manner in which these charged particles arrive and move in the upper levels of our atmosphere. Perhaps the most striking part of this phenomenon is the streaky or discontinuous structure of most aurora. It would appear as if, instead of a steady shower of charged particles, they arrived in clouds, these clouds being eventually drawn out into streaks or lines roughly parallel with the lines of force in the earth's magnetic field. This is exactly what we would expect of a charged particle in rapid motion, when it encountered a magnetic field. It would tend to spiral its way down a magnetic line of force, at least until it reached and penetrated the upper level of the atmosphere, whereupon it would begin to spend its energy in ionization, with light as one of the less objectionable by-products.

The writer is well aware that a theory is at the most tentative, and under suspicion of unsoundness, until it yields confident forecasts. Until such time as it may have been proven by this acid test, it is unwise to overburden a theory by speculative exterpolation. However, even at this risk, as well as the graver one of diversion from the subject matter of this paper, the writer cannot refrain from a speculation as to the possible connection of the auroral structure, above outlined, with certain facts as to the difference between east-and-west, and north-and-south radio communication. If, as would certainly appear from the visible aurora, east-and-west bound waves encountered a celestial hurdle, while north-and-south waves ran peacefully between the ionized filaments, the ease of north-and-south, and the comparative difficulty of east-and-west radio communication would be readily explained. One might go a step further, and attribute, at least at times, a sufficiently periodic spacing to these ionized clouds for the formation of a gigantic reflection grating, and thereby account for de Forest's[3] observation of the easy transmission of certain wave-lengths, and the high absorption of others. Certainly the visible aurora frequently has a very regular spacing, and it might be interesting to ascertain if it be of the right order to account for these effects.

But for the purposes of this paper it will be sufficient to assume only that the charged particles enter our atmosphere in a discontinuous manner. Charged clouds will be formed at some high level, probably drawn out so that their greatest length lies approximately north-and-south and, in a manner similar to that observed in a thunder-storm, there will be equalizing discharges between these clouds. At first, because of the low dielectric strength of the atmosphere at the higher levels, such discharges will occur at small potential differences, and hence will tend to be of small individual intensity but of frequent occurrence. As the charges sink toward the earth, they will reach denser air, and there the discharges will become of greater individual intensity but less frequent.

We recognize to-day two important, that is to say bothersome, varieties of static. One of these is the "click," consisting of rather widely separated but very strong disturbances, reminiscent of the discharges of that aptly-named apparatus, the static machine. The other is the "grinder," which runs the gamut of noise from a grating or grinding sound thru something resembling handfuls of gravel thrown against a window, with perhaps slushy or even hissing sounds at the other end of the scale. The clicks are most probably relatively low level discharges. This is supported by the fact that local thunder-storms give rise to clicks which can sometimes be identified with visible discharges between cloud and cloud, or cloud and ground. Grinders, however, would seem to have two possible origins. In the first place, they might arise from very high level discharges of considerable tho irregular frequency. The writer has succeeded in imitating, with a disconcerting fidelity, grinding static by the simple process of discharging a small condenser thru a series circuit of a wet string and a vacuum tube, replenishing the condenser from an influence machine. Grinding static might also arise from the summation of many distant clicks. There is some experimental evidence of the existence of these two varieties, either singly or mixed, and to this a later reference will be made. For the present, our principal interest is in the manner in which the static waves arrive at the receiving station. If we assume a more or less uniform distribution of discharges in the upper atmosphere, the condition of affairs with respect to the receiving station might be pictured somewhat as in Figure 1.

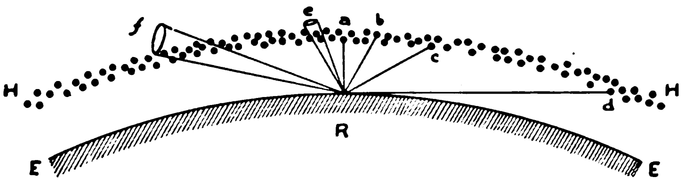


FIGURE 1

Here a discharge stratum, *H-H*, forms a portion of a sphere concentric with the earth's surface *E-E*. This stratum is not only above the receiving point *R*, but envelops this point in all directions, even to the horizon. It is at once apparent that individual discharges, as at *a*, *b*, and *c*, will set up waves or pulses which will arrive at the receiving station with intensities inversely proportional to their distances *a-R*, *b-R*, and *c-R*. This may be approximately expressed by saying that the intensities will vary as the sine of the vertical angle, save near the horizon. To state the law exactly would obviously require a knowledge of the height of the stratum *H-H* which we do not yet possess. However, it is apparent that the intensity on the horizon, as at *d*, is far from being zero, and is most probably between five and ten per cent of the zenith intensity. The maximum static intensity, tho by no means confined to the zenith, is evidently distinctly *above* the receiving point *R*. At first thought it would seem that the total amount of disturbance produced by static would be represented by the product of the individual intensity by the frequency. Inasmuch as the number of individual discharges included in a given small solid angle near the zenith, as at *e*, is much smaller than the number included in the same solid angle taken nearer the horizon, as at *f*, varying in fact approximately as the square of the cosine of the vertical angle, it would seem that the total effect should *increase* as we pass from the zenith to the horizon. In general, the writer finds this to be the fact, altho it not infrequently happens, particularly in the forenoon, that light static comes down from above, with very little from the horizon. This may be explained by considering such horizontal static as consisting of weak discharges at great distances from the receiver, and of so great frequency (because of the summation of so many individual discharges) as to blend into a distinctly "soft" disturbance. Bearing in mind that static acts upon a receiving circuit by pure impact, a not inapt analogy is that of a bell set into vibration by irregularly spaced impacts. For the same amount of energy in a given time, a bell will give the greatest response when the taps are individually strong and fairly well separated in time, rather than when the taps are weak but frequent. Furthermore, at least until such time as all reception is photographic, the psychology of the receiving operator must be borne in mind. There is doubtless an illy defined, but none the less existent frequency of maximum disturbance. For example, the writer finds irregularly spaced disturbances most troublesome when their mean frequency is about twice that of the signal element.

If the discharging clouds are drawn out in a north-and-south direction we should expect that the waves arising from such discharges would show a certain uniformity in their planes of polarization. Altho it would be absurd to think of the discharging clouds as accurately oriental linear oscillators, nevertheless there should be a certain preponderance in that direction.

The writer's theory of static might be summed up as follows: According to the assumed character and location of the discharging masses, static waves may be expected to come in on the receiving stations from all angles in altitude and azimuth, with a maximum individual intensity from above, and a maximum frequency in the neighborhood of the horizon. The total disturbing effect will in general come in from points nearer the horizon than the zenith, altho the altitude of the ring of maximum disturbance is indefinite. The individual planes of polarization may take all possible angles, but with a preponderance in a direction determined by the earth's magnetic field. Finally it would seem that the most probable static wave form was a single highly damped pulse, because of the low conductivity of the discharging masses.

According to the above theory, the problem of the elimination of static and the preservation of the signal consists simply of the reception of a wave-train originating at a definite point, and the exclusion of static pulses originating at all points in altitude and azimuth. In a broad sense, there are two solutions for this problem, one involving a separation of signal and static based upon their difference in wave form, and the other a separation based simply on sharply directional reception. It is quite clear that if we could restrict reception to a small solid angle including the distant station, there would be relatively little static included. To use an optical analogy, the problem is essentially like that of receiving monochromatic light signals in full daylight. A spectroscope or a filter screen would be one solution, and directive reception as by a telescope another, but best of all would be a combination of the two. In present day radio communication we have already highly developed the spectroscopic or filter separation of signal from static, as by sharply resonant circuits and beat reception. But altho we have had two practicable types of sharply directional receiving circuits for the past twelve years or more, very little use of these appears to have been made until quite recently.

In a paper entitled "Absorption and Reflection of Electrical Waves," read before the New England Wireless Society on December 7, 1912, the writer, after discussing the probable relation between ionization conditions in the upper atmosphere and electrical wave transmission, went into some detail of his early experiments to determine the nature of static. At his first experimental station, at Blue Hill Observatory, Milton, Massachusetts,[5] he conducted a long series of tests which apparently showed a relation between the wind velocity and the number of disturbances in a given time, which led to the conclusion that effect was a local one, possibly consisting of an actual discharge to the receiving aerial from electrified masses of air moving past the antenna wire. This theory was rudely upset by his further work in 1901, when comparison of tapes at Galilee and Brielle, New Jersey, about twenty miles (32 kilometers) apart, showed the same dots and dashes from static, day after day. It was clear from this that some static, at least, came from discharges at considerable distances from the receiving aerial. He then said:

"A little later, I surrounded a receiving aerial by a F raday cage[6] constructed with loading inductances to avoid shielding the aerial inside from the desired signal. This seemed to eliminate a small portion of the static—probably that portion due to direct discharge to the wire from the atmosphere—but had no observable effect upon the distant variety."

At this point, the writer wishes to call attention to the careless manner in which members of this Institute refer to the Faraday cage antenna shield as a "Dieckmann cage." Dieckmann, in 1912,[7] first published his description of an aperiodic shield or cage around an aerial, so designed as to permit the passage of signal waves, but to be opaque to static pulses. The writer has published, in a patent[6] filed six years before Dieckmann's date, this exact thing. Reference to Figure 2 will show the antenna wires *A*, *A*, surrounded by a cage, sectioned off into short elements which are connected together by impedances and grounded thru other impedances so that an aperiodic shield results. While he is inclined to agree with Weagant[12] that the device is of small utility as against the most bothersome varieties of static, it seems only fair that he, and not Dieckmann, should receive whatever credit is due for this invention. While the

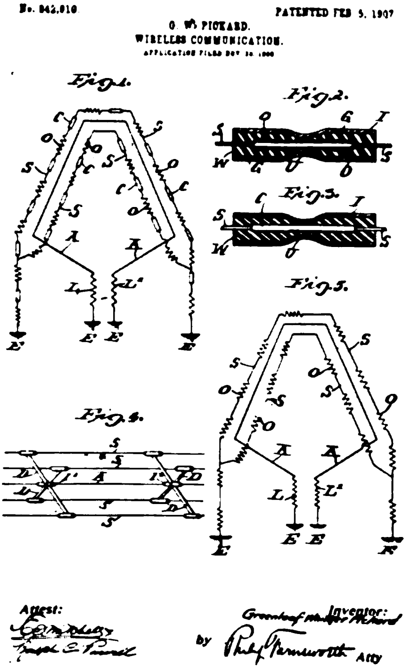


FIGURE 2

shielded antenna is open to attack from clicks and grinders, it is a complete solution of the hissing variety of static, due to a direct discharge to the aerial. Where a series condenser is employed in the antenna circuit, this is very apt to charge up to the limit of its dielectric strength, and then spill over, giving an excellent imitation of pistol shot static. This occurs most markedly during snow-storms; and, on shipboard, where the aerial is slung immediately over the funnels, the vessel manufactures this type of static in abundance, and delivers it effectively to the waiting antenna. Shielding is of distinct value in such cases.

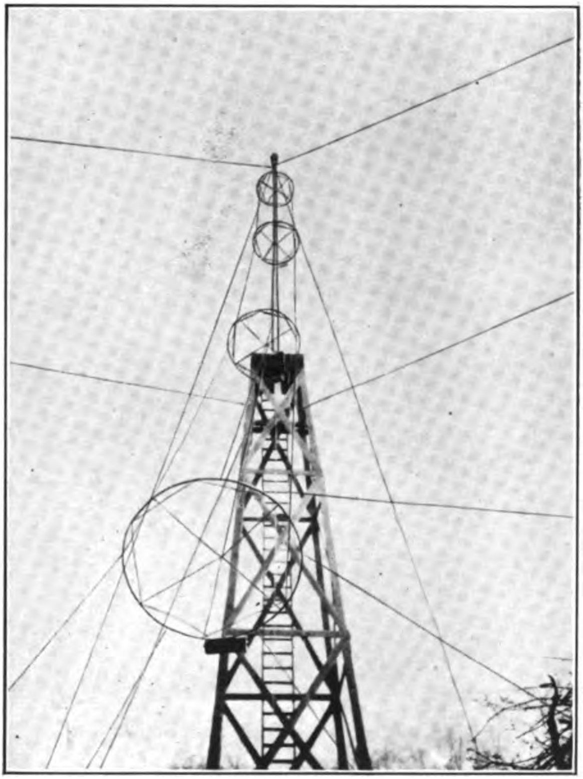


FIGURE 3—Pre-Dieckmann Shield

Continuing his paper before the New England Wireless Society, the writer then describes his early work with directive aerials, particularly with the ungrounded coil or loop aerial.

Again, as with the Dieckmann cage, the writer must pause to correct a heresy. Altho it has been fashionable of late to attribute this invention to Messrs. Bellini and Tosi, it should be noted that these gentlemen first published an account of their work on November 14, 1907[13]. The writer began his development of the coil or loop aerial in 1902, and published an account of his work on June 15, 1907[13], almost exactly five months prior to Messrs. Bellini and Tosi. It is of interest in this connection to note that the writer, on September 13, 1907, demonstrated one of his coil aerials at Dorchester, Mass., to Lieutenant-Commander S. S. Robinson, detailed by the United States Navy for this purpose. This particular coil aerial is shown in Figure 5, and consisted of two turns of wire forming a square about nine meters (27.45 feet) on a side. Altho but two turns are shown,

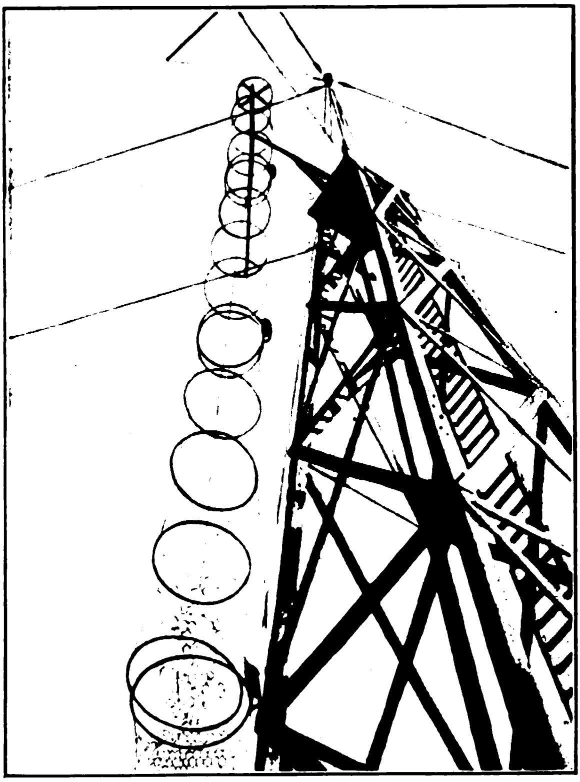


FIGURE 4—Pre-Dieckmann Shield

from one to ten turns were actually employed, depending of course upon the wave-length. Exceedingly accurate bearings of a number of distant stations were taken, some over sixty kilometers (38 miles) away from the loop, with a maximum error not greater than 2 degrees of arc.

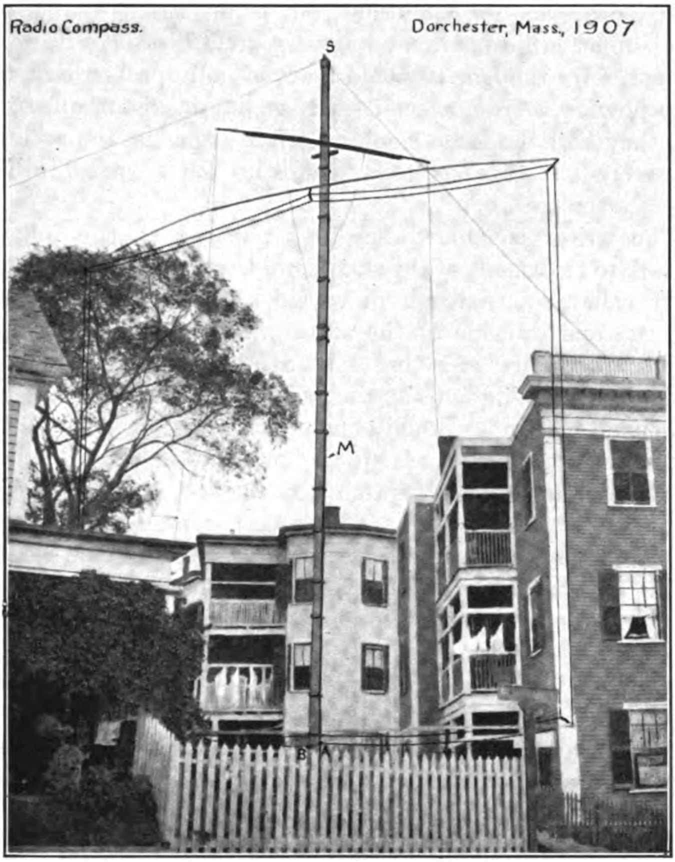


FIGURE 5

Further on in the 1912 paper, the writer said:

"But the worst summer static appears to be at a very considerable distance from the receiver, and I believe it originates at some of the higher levels already discussed in this paper; probably many miles over our heads. I also believe that static is simply a discharge from one high level cloud of ions to another similar cloud, very like what we see from cloud to cloud in a thunder-shower; that is, an equalizing discharge. Like lightning, static appears to be non-oscillatory, as may be readily found by experiment."

The writer does not claim novelty for the "static from above" theory. In fact, Airy[4] said in 1911:

"It was concluded that most of the disturbances were not due to local weather conditions, but to discharges taking place in the upper atmosphere, and at very great distances from both stations. He thought it would be eventually proved that these atmospherics were connected with magnetic disturbances, and that they had the same common origin—that is, the arrival of negatively charged electrons from the outer space into the earth's atmosphere."

The writer concluded his 1912 paper with the following up-to-date treatment of the static problem:

"I believe, as a result of considerable experimental work, that the real solution of the static problem lies in the use of sharply directional receiving antenna. Provided only that the static does not originate at the same point as the signal, good directional reception should eliminate most of it."

As already stated, there are two practical methods of obtaining sharply directional reception. In the first method, originally suggested by Elihu Thomson[8] in 1899, patented in England by S. G. Brown in the same year[9], more fully disclosed and patented in this country by Stone in 1902[10], elaborately worked out for two, three, and four collecting circuits by Braun in 1906,[11] and more recently used by Weagant[12], two or more separate collectors are employed, spaced apart a material fraction of a wave-length in the line of propagation, as shown in Figure 6. These collectors act upon a common secondary circuit in such wise that when the collectors are simultaneously affected by any disturbance, the current set up therein will arrive in phase at the center of the system, and there, by opposed windings, be placed 180° apart in phase in their effect upon the secondary. Because of this opposition, such currents will not affect the detector, and no signal will be produced. On the other hand, the signal wave, passing the collectors successively, will set up currents therein which will arrive at the center of the system in an out-of-phase relation, and such signal currents will add in their effect upon the secondary and affect the detector. This is very well shown in Figure 7, and clearly explained in the specification of the Stone patent of which Figure 7 is the drawing. The writer, in 1905, devised

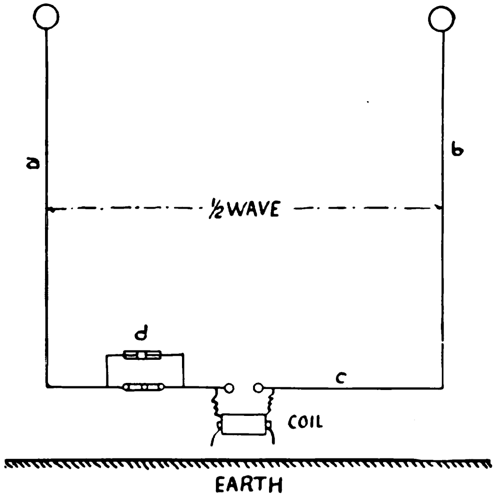


FIGURE 6

a system in which two aerials, tuned to the distant station and separated on centers by some thousands of meters (about a mile),

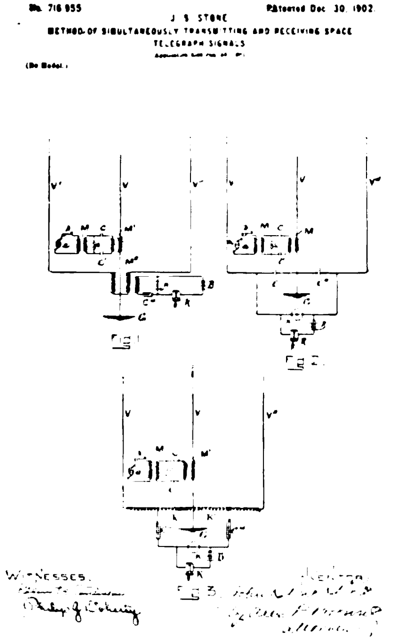

No. 716,955
Patented Dec. 30, 1902.
J. S. STONE.
METHOD OF SIMULTANEOUSLY TRANSMITTING AND RECEIVING SPACE TELEGRAPH SIGNALS.
(No Model.)
V
M
C
B
K
G
WITNESSES.

FIGURE 7

were connected to a common secondary by way of phase adjusting means, so that the currents were added in the secondary in such phase as to produce the maximum signal. This method was first published in a patent filed in 1907,[13] and is shown in Figure 8. The particular case discussed in this publication involved two large loop or coil aerials, arranged in an east-and-west plane for the purpose of trans-atlantic reception. These loop aerials consisted of long, low rectangles, and were tuned to the desired wave-length by inserted series capacity and inductance. With this system it is possible to so adjust the phase relations of the currents arriving at the secondary that interference or static arriving from any particular direction is cancelled out in the secondary, and does not affect the detector. A further advantage of this system, which has recently been pointed out by Weagant[14], is the fact that it has materially greater or sharper directional reception than any single type of collector. This is obvious when we consider the fact that the loop aerials M^2 and M^4 have simple cosine reception curves in the horizontal plane,

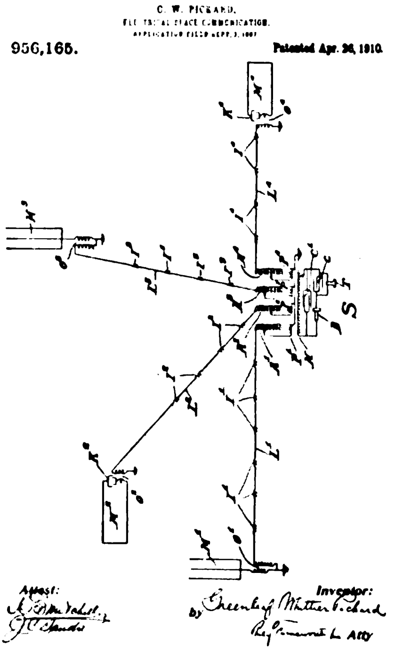


FIGURE 8

and that a pair of open or non-directive aerials, separated as are M^2 and M^4, in Figure 8, by a considerable fraction of a wavelength, with currents combined in a common secondary, have also a cosine curve characteristic. The combination of two loop aerials in the manner shown in Figure 8 has therefore a cosine square reception curve.

This same publication also states that combined loop and open aerials may be used at such points as M^2 and M^4. It is of interest to note the reception curve of this system, when one, two and three separate collectors of this form are employed. In Figure 8-A is shown in dotted line the reception curve of one collector. This is a cardioid, or heart-shaped figure, with maximum reception in the direction *O-A*. With two such collectors, the reception curve is shown in dot-and-dash line, and is markedly more sharply directional than the curve for one collector. A still further improvement results when three collectors are employed, this being shown in full line.

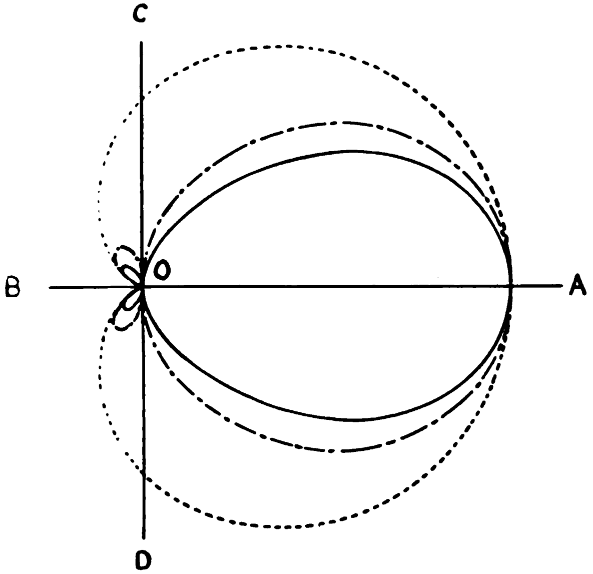


FIGURE 8-A

The second method of directional reception, devised by the writer[21], does not involve, even for long wave reception, such geographical dimensions as the systems shown in Figure 8. This second method consists simply of a closed tuned circuit or coil aerial, of dimensions small as compared with a wave-length. In its most effective form, consisting of a coil aerial combined with a so-called "open" aerial, this method gives true unilateral reception. The coil may be of quite small dimensions, particularly when it is employed simply as a radio compass. In Figure 9 is shown the first portable radio compass, which was employed by the writer in 1907-1908[16] for mapping out the wave-front around a transmitting station. This radio compass consisted of a three-turn loop, one meter (3.05 feet) in diameter, shunted by a variable air condenser and a crystal detector.

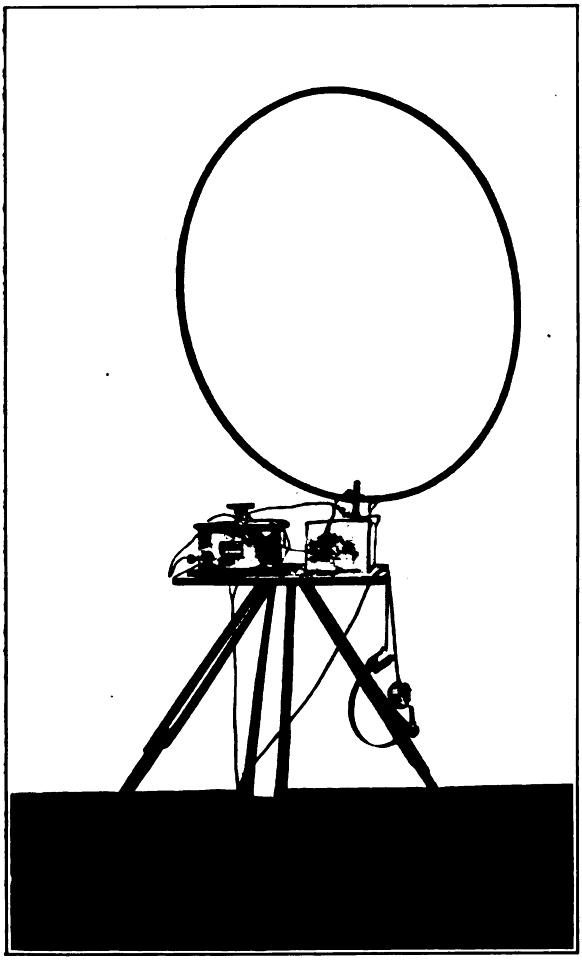


FIGURE 9

Heretofore, the reception characteristics of aerials have been studied principally, if not exclusively, in the horizontal plane. The writer is not aware of any publication dealing with the three-dimensional characteristics of aerials, altho for quite some time past, points of wave origin other than the horizon have been common, for example, static, and more recently transmitters on aircraft. Just why radio engineers have elected to live in a two-dimensional world is rather puzzling, and if this paper has no other effect than to add another dimension to their life the writer will feel fully repaid for his labor.

In Figure 10 is shown the now familiar reception curve of a vertical loop or coil antenna in the horizontal plane, consisting of a cosine curve drawn in polar co-ordinates. This is an ideal curve; actually a coil aerial gives a more or less distorted figure-of-eight, sometimes tending to an hour-glass shape, and usually more or less unsymmetrical. The reasons for this distortion have only recently been worked out in detail, and some of the more important causes of distortion will be explained later in this paper.

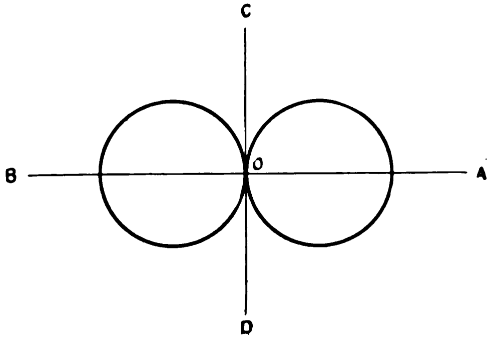


Figure 10

In a vertical plane, including the plane of the loop itself, reception is obviously symmetrical, and the reception curve in this plane is a circle, as shown in Figure 11. If the loop is near the ground, the lower half of the reception curve is normally without interest, because signals do not often originate from under foot, and hence is shown in dotted line.

If, however, a loop is placed aloft, as in aircraft work, the useful reception curve in this vertical plane is a full circle. Finally, the complete three-dimensional reception surface of the

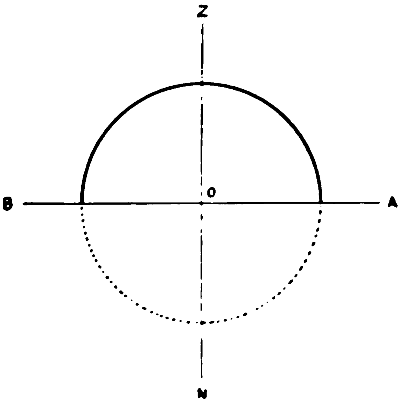


FIGURE 11

loop is a torus. This torus is shown in Figure 12, and differs from the conventional anchor ring or doughnut in that it has no hole in the center, this peculiarity making its representation somewhat difficult. The significance of this last figure is obvious —the loop receives more or less from all points in space, excepting only along a line passing normally thru its center. The further

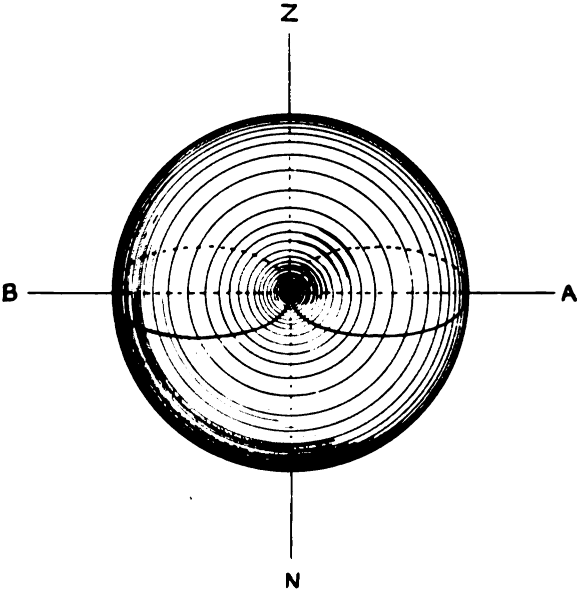


FIGURE 12

meaning of this figure is that maximum reception occurs only in all directions in the plane including the loop, and the significance of this fact is perhaps best brought out by Figure 13. This figure represents a plan of the hemisphere over the loop, just as if one were looking up at the zenith, and the shading indicates the amount of reception from different points in this hemisphere by a loop placed in an east-and-west vertical plane. The zone of maximum reception is a belt passing across the sky from east to west, reception decreasing to zero on the horizon at points north and south of the loop.

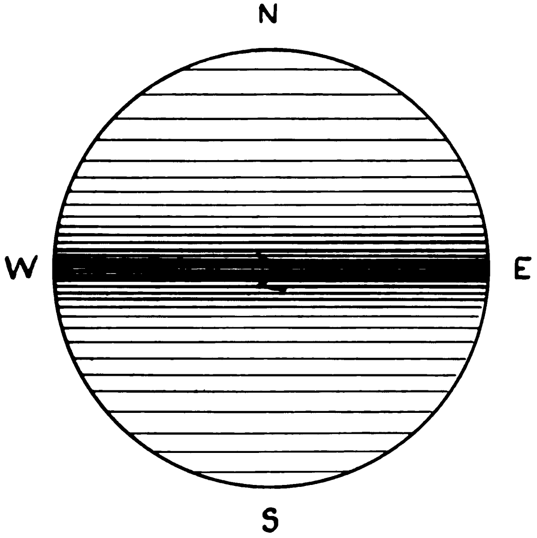


FIGURE 13

A so-called open antenna consisting of a vertical wire, short as compared with a wave-length, has in the horizontal plane a simple circle diagram, that is, it receives equally well from all points on its horizon. In any vertical plane passing thru the antenna, a quite different reception curve results, as reception is a maximum on the horizon, and decreases to zero at the zenith. The complete reception in any vertical plane passing thru the antenna is shown in Figure 14, and is, of course, the familiar cosine curve. The portion below the horizon is shown dotted in the figure; if the open antenna consisted of an ungrounded linear oscillator considerably removed from the earth, the lower portion of the curve would interest us. Actually, just as with the loop or coil aerial, and for similar reasons, this ideal curve usually suffers considerable distortion, partly because actual vertical aerials have an appreciable horizontal length in most cases, and partly because of unsymmetrical or electrically warped surroundings. The complete three-dimensional reception surface of the open antenna is therefore just like Figure 12 laid down flat, or, more exactly, laid down flat and half embedded in the ground if the open antenna is on the earth's surface. It will be seen from this that the loop and the open antenna have identical reception surfaces, but in planes 90° apart. Similarly to the loop, Figure 15 shows a plan of the hemisphere over an open antenna. In this

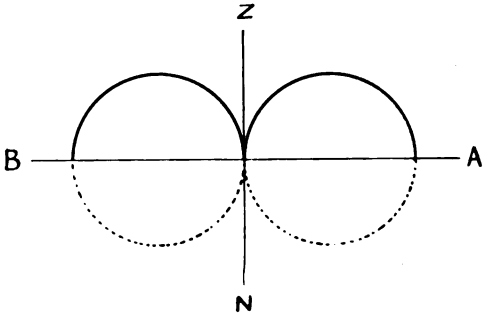


FIGURE 14

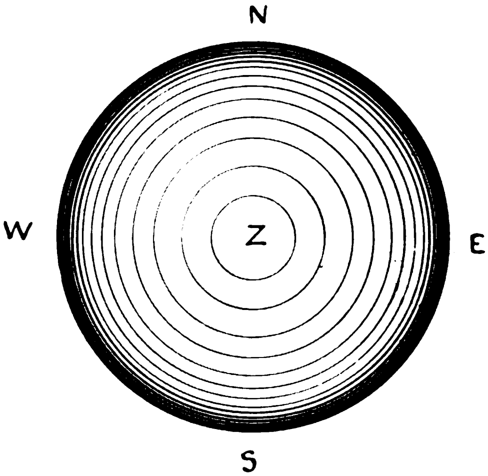


FIGURE 15

figure, the reception belt rings the horizon, and reception is zero from the zenith. If static came solely from above, an open antenna would make an ideal eliminator. Equally, if static came in on the receiving point with the same intensity from all points in altitude and azimuth, it would be difficult to account for the well-known freedom of the loop from static.[20] But if we assume that in general static is at its worst near the horizon rather than the zenith, a comparison of Figures 13 and 15 will show a reason for the relative immunity of the loop.

In Figure 16 is shown the method referred to above, which combines open antenna reception with loop or coil reception. A glance at the date of this drawing—June 10, 1907—will make it obvious that the combination of loop and open reception cannot be attributed to Messrs. Bellini and Tosi, often as this has been attempted of late. In this figure is shown the coupling of a loop and an open antenna to a common secondary, by way of phase-adjusting means, so that the currents in the loop and the

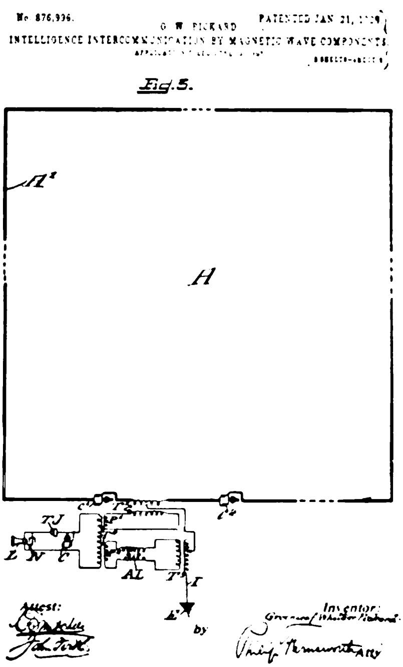


FIGURE 16

open antenna, normally 90° out of phase, are added in phase in the secondary circuit. If now, as by suitable dimensions of the loop or by coupling adjustment, or by both, the currents from the open antenna are made equal, in their effect upon the secondary, to those from the loop, the result of this addition, in a reception curve on the horizontal plane, is shown in Figure 17. Reception is at a maximum from the direction *O-A*, is zero from the direction *O-B*, and has intermediate values from other directions, the complete curve being a cardioid. Such a circuit, employed as a direction finder, gives the true bearing of the distant station, and does not, as with the simple loop, leave the direction indeterminate by 180°. In a vertical plane including the loop the reception curve is also a cardioid, as shown in Figure 18, with a maximum on the horizon at *A*, and zero on the horizon at *B*. Reception from the zenith, that is, along the line *O-Z*, being limited to the loop, has half the value of reception along the line *O-A*. In a vertical plane at right angles with the loop, that is to say, at right angles to the plane of Figure 18, the reception curve is that shown in Figure 19. In the center of this figure is shown the loop *L-L* and the open antenna *V-V*. A static source at *S-S*, in the vertical plane *C-Z-D-N*, is at an angle *A* with the vertical, and has an orientation or polarization angle *B*.

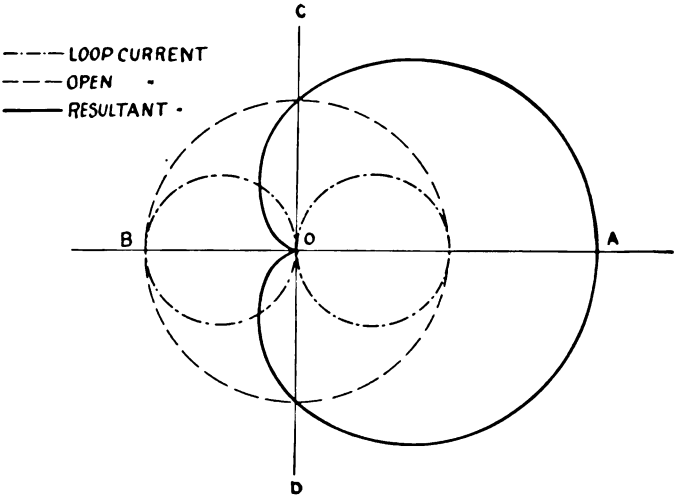


FIGURE 17

The current set up in the loop will be proportional to $\cos A \sin B$, while the current excited in the open antenna will be $\sin A \cos B$. The sum of these currents will be $\cos A \sin B + \sin A \cos B$; that is, for all possible orientations and vertical angles it will be a constant, so that the reception curve in the vertical plane normal to the loop will be, as shown in the figure, a circle.

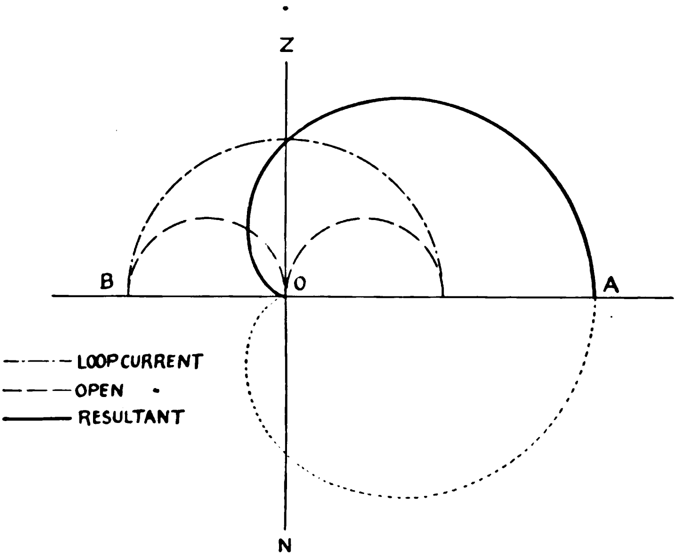


FIGURE 18

The complete three-dimensional reception surface is difficult to show in a two-dimensional figure or plan. Figure 20 shows its sections in three perpendicular planes. As will be apparent from this figure, the complete reception surface is a cardioid of revolution, which may be more simply described as an apple, with the pit at the origin O.

From a consideration of the three-dimensional unilateral reception of the combined open and loop antenna, as shown below in Figure 20, it will be obvious that static originating at points other than on the horizon in the direction of the distant transmitter will either be weakly received, or, if it happens to originate on the horizon at a point 180° from the transmitter, it will not be received at all. The circuit shown in Figure 16 is not at present, however, the best arrangement for this method of static reduction. As the effect of static on a receiving circuit is practically pure impact excitation, complex waves will be set up in the two coupled circuits—the loop and the open antenna—with the result that the just shown ideal diagrams of reception curves and surfaces are but roughly approximated. These diagrams are, of course, only true for undamped wave reception, or for static reception on substantially aperiodic circuits.

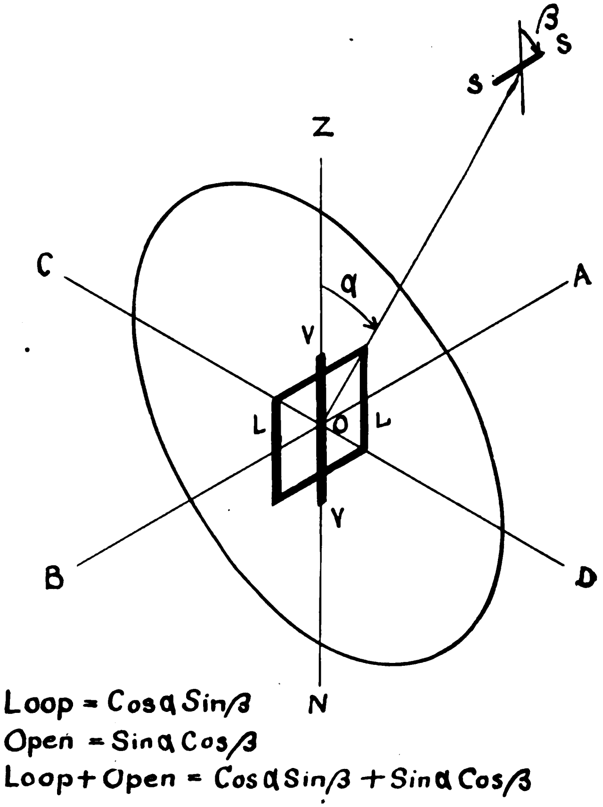


FIGURE 19

Shortly after America entered the war, the Wireless Specialty Apparatus Company equipped a radio station for Mr. Alessandro Fabbri, at Otter Cliffs, near Bar Harbor, Maine. When the station was completed, Mr Fabbri donated it to the Navy, and it was first used as a low power spark station. However, it was very shortly found that the location was an excellent one for trans-atlantic reception, and this eventually became its principal service, the station finally becoming the premier reception point in this country. The summer of 1917 having shown that altho the Otter Cliffs station had excellent signals, it also had overwhelming static on occasion, Mr. Fabbri invited the writer to install directional receiving circuits for static elimination, adapted to the requirements of trans-atlantic service. Within a few days after this request, the circuit shown in Figure 21 was installed, combining open antenna reception with loop or coil reception. This was immediately found successful in improving

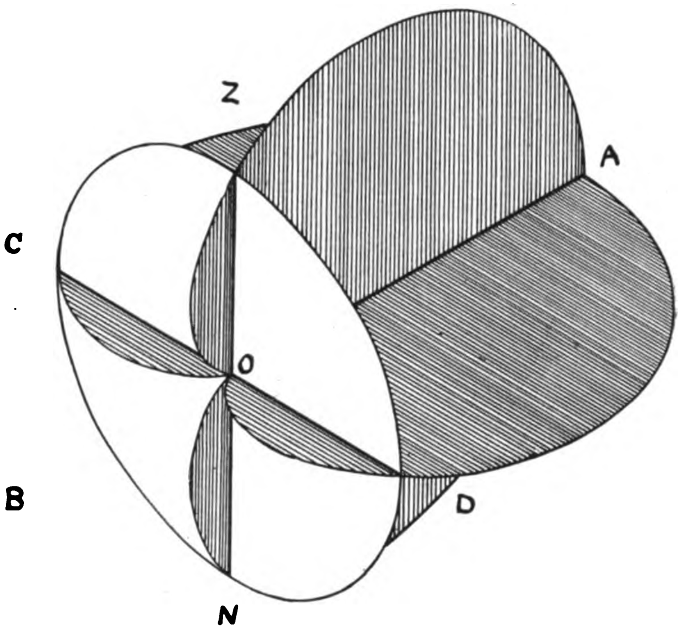


Figure 20

the signal-static ratio, and was at once placed in service. The loop A consisted of a solenoidal coil of four turns of number 16 Brown and Sharp gauge (1.3 mm. diameter) copper wire, with the turns spaced apart 30 cm. (12 inches). This loop was in the form of a long low rectangle, 30 meters (97.5 feet) long, and 6 meters (18.3 feet) high, the lower part of the loop being about 4 meters (12.2 feet) from the ground. The plane of this loop was approximately north-east by south-west, that is, it was in the great circle bearing of the more important European stations. This coil aerial was tuned to the desired wave-length by series inductance L_3 and capacity C, a fixed coil L_3 of 22 millihenrys being used in most of the work. The open antenna of this combination consisted of the conductor of the loop itself and a connection to earth at *G* by way of a coupling coil L_1 of 22 millihenrys inductance, a variable inductance L_2 for tuning this open antenna circuit, and a variable resistance *R*, and was tuned to the same wave-length as the loop. The coils L_1 and L_3 were quite closely coupled together, and coil L_3 rather more loosely coupled to the secondary or detector circuit S. This secondary circuit, as shown in Figure 21, employed the static coupled audion circuit due to Lieutenant Eaton of the United States Navy.

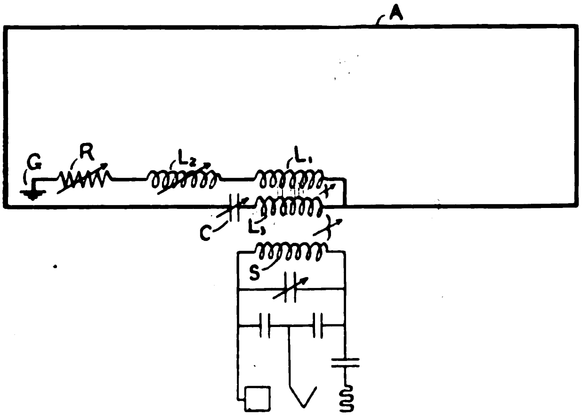


FIGURE 21

Two steps of audio frequency amplification were usually employed, in addition to the circuit shown, and the normal signal strengths from Rome, Lyons, and Carnarvon were above 1,000 times audibility. The resistance *R* was found to be an important element of the circuit, and in its absence the results were decidedly inferior. The amount of resistance required was surprisingly large, being about 3,000 ohms.

The operation of this circuit is substantially as follows: As it is a development of the circuit shown in Figure 16, and consists of a combination of a loop and an open antenna, with the currents added in phase in the secondary circuit, it has a reception surface which is essentially like the upper half of Figure 20, with a maximum reception only on the horizon in the direction of the distant transmitter. Static pulses from any direction other than that of the distant station itself are either received weakly or not at all if it happens to originate to the rear of the system.

Reaction between the open and closed circuits, with resultant complex wave formation, is prevented by the high damping of the open circuit, which with 3,000 ohms inserted is practically aperiodic. The reason why so large a resistance can be used in the open circuit and still have substantial equality of open and closed circuit current is due to the fact that a loop of the dimensions employed is much less effective than an open antenna of the same height at long wave-lengths. The currents from the open circuit, reduced to the same amplitude as those in the loop by the inserted resistance and degree of coupling, are shifted thru 90° in phase by the transfer from the open to the closed circuit by way of the coupling, between L_1 and L_3, and hence are added in phase in the loop circuit, and then transferred, by the coupling between L_3 and S, to the secondary circuit. It will be seen that the change in phase accomplished in the original circuit of Figure 16 by a phase adjuster in the form of a variable artificial line, is here done by the simple expedient of coupling together tuned circuits.

The performance of this circuit may be judged from a typical set of audibility meter readings taken by the writer, which are tabulated below:

Time A.M.	Station	Loop alone		Loop+Open	
		Signal	Static	Signal	Static
11.25	Nauen	2,500	800	400	60
11.35	"	5,000	1,500	3,000	800
11.45	"	3,000	1,000	400	120
11.50	"	5,000	1,500	——	——
12.00	"	3,000	1,500	——	——
P.M.					
2.35	"	1,000	1,000	250	60
2.40	"	——	——	1,000	200
3.10	Carnarvon	1,000	4,000	600	100
3.15	"	1,500	5,000	600	160
3.25	"	800	5,000	200	80
3.30	Nauen	1,000	5,000	250	80
3.35	"	1,000	4,000	600	160
3.37	Carnarvon	1.000	5,000	600	160
3.41	"	1,000	4,000	500	160
3.55	Nauen	1,000	5,000	300	80
4.00	"	600	2,000	300	60

The static intensities in the above table were taken in each instance over a period of five seconds, and no attention was paid to the clicks, which were of several times the intensity of the grinders. It will be noted that during the forenoon, with strong signals and weak static, the system had little effect. But during the afternoon, with weaker signals and much stronger static, the improvement in the ratio is rather striking. As a matter of fact, the circuit of Figure 21, during periods of severe static, enabled the solid, uninterrupted copying of stations which would on the ordinary circuits be so broken up as to be unreadable.

Continued use of this circuit at the Otter Cliffs station developed the fact that the earth connection of the open antenna was not attached symmetrically to the aerial, the result being that a certain amount of current in the open circuit was flowing directly thru the loop and coil L_3, of course 90° out of phase with the loop current. This tended to decrease the sharpness of the unilateral reception, and hence the degree of static elimination. As this effect is often present in the simple loop or coil aerial when used for direction finding, it may not be out of place to explain it briefly here.

Considering only the horizontal plane reception of a simple loop aerial, with a certain unbalanced capacity to ground (either by way of unbalanced leads or unsymmetrical construction of the loop itself), it is obvious that the system will act as a combination of loop and open antenna, with the currents added together, not in phase but in quadrature. As a result of this, the currents will add together irrespective of their direction, with the result that the ideal figure-of-eight reception curve degenerates into an hour-glass shaped figure, the breadth of the neck depending upon the amount of open antenna current. This degeneration is shown in four steps of increasing open antenna current in Figure 22. The remedy for this trouble is a simple one, and consists in balancing the loop to ground, as by a three-plate variable condenser, in the manner shown in Figure 23. By turning this loop thru 180° in its own plane, so that the terminals of the loop come out on top, a material advantage results in direction finding. As the points of highest potential are thus removed as far as possible from the earth, the loop is much less affected by surrounding grounded objects. Of course, this method is at its best with a single turn loop, altho improving a two-turn loop.

A simple loop may also suffer from open antenna current which is neither in phase with the loop current, nor in quadra-

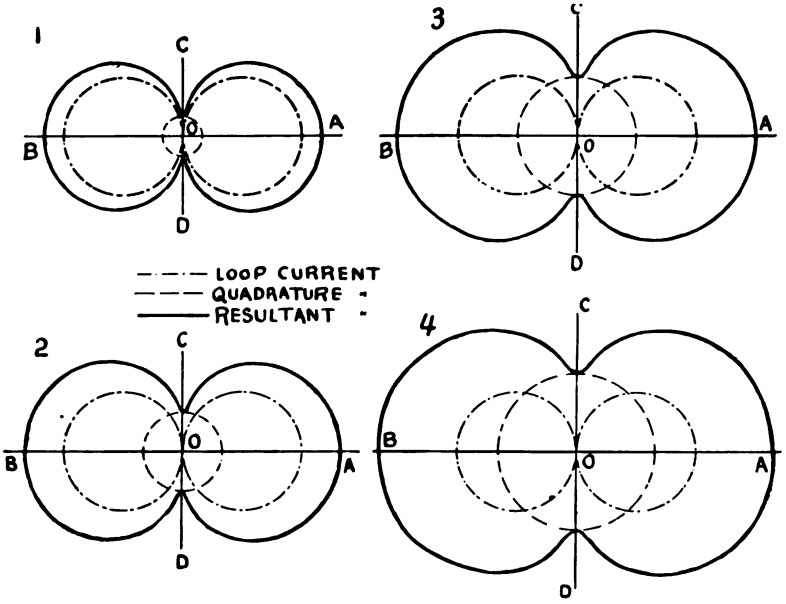


FIGURE 22

ture therewith. This sometimes happens in short wave working with loops of considerable inductance, that is, of considerable

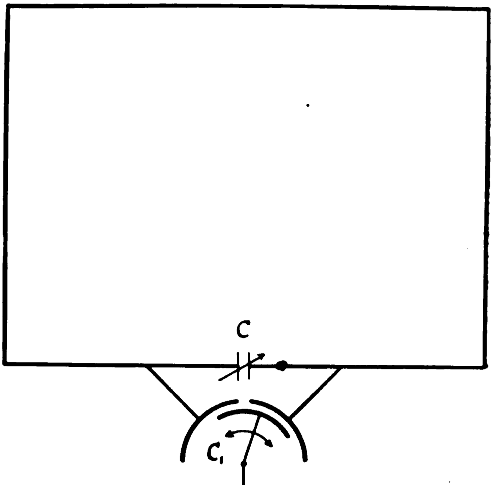


FIGURE 23

conductor length. It also happens with the circuit of Figure 21, because the two open antenna effects, one in phase and the other in quadrature, may be considered as one current. For graphical analysis, it is simpler to keep these currents separate, and this is done in Figure 24, which illustrates the evil effect of out-of-phase current. This is also an unbalance effect, and may be eliminated by suitable construction or by the three-plate condenser arrangement of Figure 23.

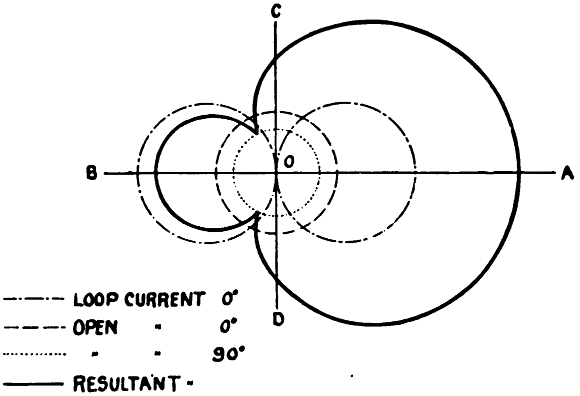


Figure 24

A more bothersome effect in simple loop working arises from a displacement current across the turns of the coil aerial, added to an open antenna current resulting from unbalance. Inasmuch as the displacement current across the turns of the coil gives a cosine characteristic like that of the loop itself, but displaced by 90°, the resultant characteristic of the system suffers distortion which is particularly noticeable at the minimum points at right angles to the loop. The writer had intended to treat this at some extent, but the interesting paper of Captain Blatterman[18], which issued after this paper was written, makes this unnecessary. It is sufficient to say that, with the exception of the displacement current effect (which can be obviated by using a pancake rather than a solenoidal coil aerial) proper balancing of the loop, either by construction or by the three-plate condenser, is all that is required.

An interesting modification of the circuit of Figure 21, due to a suggestion by Mr. J. A. Proctor, was tried out at Otter

Cliffs, and found to reduce the unbalanced open antenna effect. This circuit is shown in Figure 25, and differs only from the circuit of Figure 21 in that the ground connection from the right-hand end of coil L_3 is taken by way of a second loop, A_1, closely coupled with the main loop A. Unbalanced current in the loop A is opposed by current in the loop A_1, with the result that there is a certain cancellation of effect, and a sharpening of the directional action.

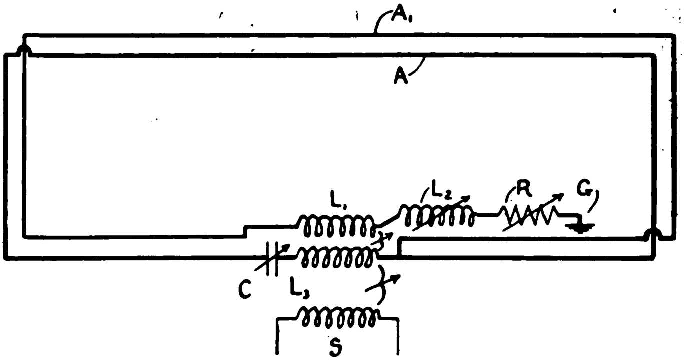


FIGURE 25

But the simplest solution, and the one that has given the best result from the combination of vertical open antenna and loop, is shown in Figure 26. This differs only from Figure 21 in that a separate open antenna is used, and the loop circuit remains entirely insulated. A typical performance of this

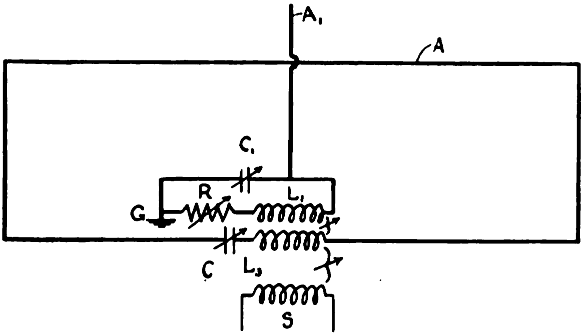


FIGURE 26

circuit (which had a loop of approximately the same dimensions as that of Figure 21) is given below:

Time P.M.	Station	Open alone		Loop alone		Loop+Open	
		Signal	Static	Signal	Static	Signal	Static
1.45	Nauen	2,000	4,000	300	600	300	60
1.54	Hanover	400	1,500	100	500	200	80
2.38	“	400	1,500	150	500	300	100
2.44	Nauen	800	6,000	200	500	300	80
3.10	“	200	800	300	800	400	80
3.18	Lyons	100	1,000	150	500	150	40
3.27	“	200	2,000	150	1,000	200	60
3.52	“	300	1,500	100	500	150	50
7.14	Carnarvon	300	1,500	150	600	150	80
7.28	“	400	4,000	300	1,000	400	100
7.40	“	400	3,000	200	800	300	100
8.26	Lyons	400	1,500	150	500	300	100
8.57	“	400	1,500	200	1,150	300	150
9.11	Carnarvon	100	800	100	800	150	60
9.20	Lyons	600	8,000	200	1,500	150	100
9.32	Carnarvon	?	8,000	?	1,000	150	100
9.37	Lyons	300	1,500	100	900	100	60

One of the striking things in the above tabulation is Carnarvon at 9.32 p. m. Neither on open antenna nor on loop, taken singly, could sufficient signal be heard to measure on the audibility meter. But on the combined loop and open antenna the signal was well above the static in intensity.

The adjustment of the combined loop and open antenna, in either of the forms shown, is quite critical, the best unilateral reception being when the loop and open antenna currents are equal. In Figure 27 is shown the result of combining different ratios of open and loop current. The uppermost reception curve is the horizontal plane characteristic of an open antenna, with no loop current. The second curve represents equal amounts of open and loop current, and is a cardioid.

The third figure represents the addition of two parts of loop current to one of open, while the fourth shows pure loop reception. It will be seen that the vector *O-A*, in the direction of the distant transmitter, increases uniformly as loop current is added to open current, until the ratio is one to one. Any further change in this ratio, increasing the proportion of loop current, decreased the vector *O-A*, the reception being only at a maximum when the ratio is unity. Similarly, the amount of reception from the direction *O-B* decreases steadily as loop current is added to open, being zero when the ratio is unity. Any further change, increasing the ratio of loop to open, results in increasing the amount of reception from *O-B*.

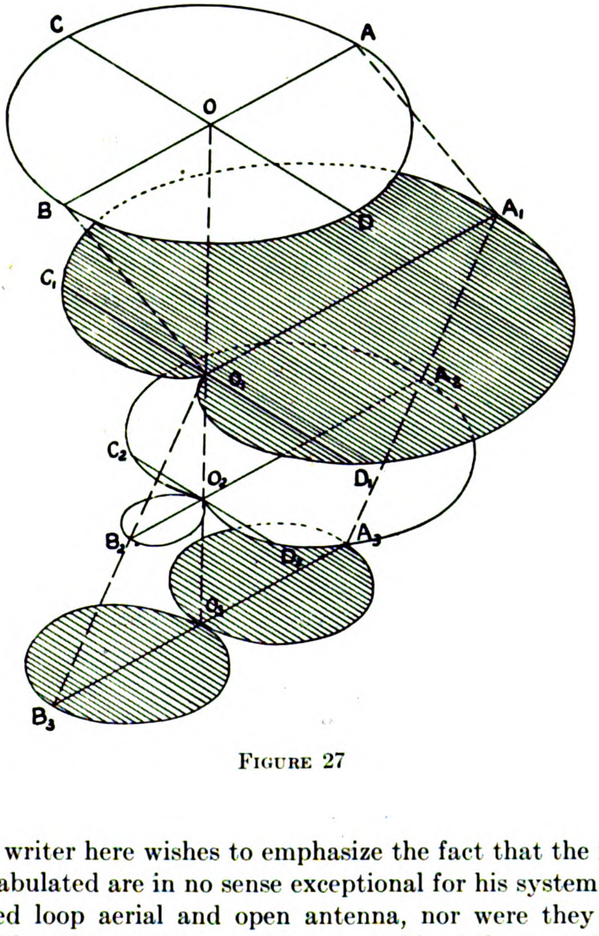


FIGURE 27

The writer here wishes to emphasize the fact that the results above tabulated are in no sense exceptional for his system of the combined loop aerial and open antenna, nor were they taken under "freak" conditions. These combined loop and open circuits have been in daily operation, summer and winter, over a period of nearly two years at Otter Cliffs, and have proven consistent performers. During the summer of 1918 the Otter Cliffs station circuits gave unbroken copying of the European stations at times when all other Atlantic coast receiving points were helpless. In fact, altho a large amount of interesting experimental work was done at various points on the Atlantic coast, it may now be definitely said that the above-described circuit, in the installation at Otter Cliffs, was the only static eliminator that played any part in our war use of trans-atlantic radio communication.

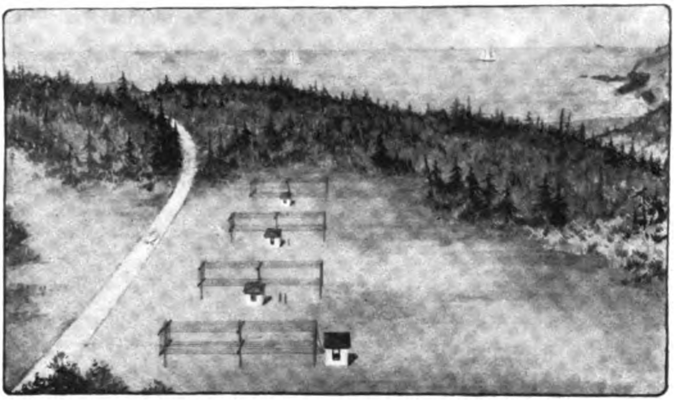
An interesting and as yet unexplained matter is the marked predominance, at least on the northern Atlantic coast, of static originating on the western or south-western horizon. In all probability, this direction of origin accounts for some of the elimination observed on the above circuits. The writer also has a certain amount of evidence which tends to show that this effect is confined to the immediate neighborhood of the coast, and that it vanishes a short distance inland.

As a result of the writer's study of the three-dimensional characteristics of various aerial combinations many novel and useful arrangements have resulted. While a detailed showing of these arrangements will be given in a later paper, a brief description of one of them may be of interest here. Examination of Figure 20 will show that if this figure is turned thru 90°, so that the vertical plane *C-Z-D-N* becomes the horizontal plane, a very favorable reception surface results. No energy would be received from the zenith, and but very little from the rest of the hemisphere, until the horizon was reached.

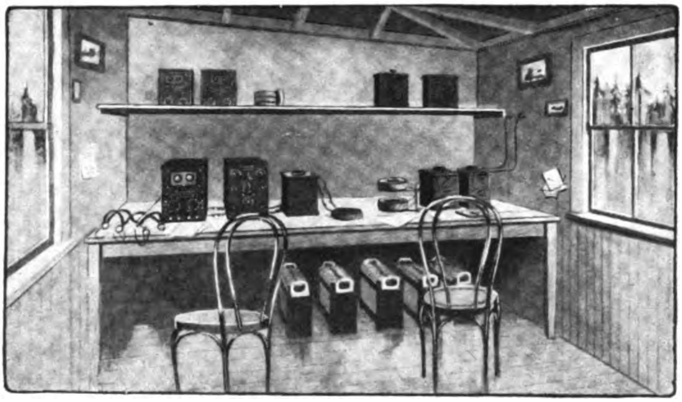
On the horizon, from static sources having all possible planes of polarization, there would be reception from all directions, but for signals from distant transmitters the predominant reception would be from the loop, so that signal reception would be directional.

In order to turn the reception surface of a circuit thru 90° one has only to turn the circuit itself thru this angle. In Figure 28 is shown the circuit of Figure 26, turned thru 90° in the plane of the loop. It is best to remove this circuit somewhat from the earth's surface, in order to prevent distortion of the open antenna characteristics. Ordinarily, if the lower part of the loop is elevated some five or six meters from the ground there will be no appreciable distortion.

In this connection it must be remembered that, over land at least, wave-fronts are not vertical, but are tilted forward in the direction of propagation. For example, at Otter Cliffs the writer found that waves arriving from European stations were inclined over 15° from the vertical. This tilting of the wave-front in and over poorly conducting soil is responsible for many curious effects; among others, it is the explanation of the action of the so-called underground aerials. When a tilted wave-front passes the circuit of Figure 28, it is equivalent to tilting the horizon of this figure, with all that that implies as to the reception surface. Returning to Figure 20, it will be seen that if the plane *C-Z-D-N* be rotated slightly about *C-D* as an axis, one half of the reception surface will be more or less submerged, so that



Static Eliminator at Otter Cliffs with Four Stations for Simultaneous Reception of Trans-atlantic Schedules



Interior of Static Eliminator Station

reception will be unilateral. In this manner, by swinging the circuit of Figure 28 thru a slight angle, it is possible to suppress practically entirely one side of the reception, and with the normal south-western static and trans-atlantic reception the circuit is free from disturbance.

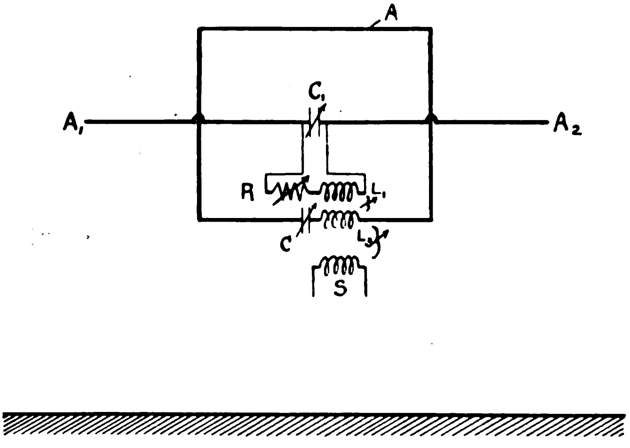


FIGURE 28

The writer gives below a typical extract from the performance log of this circuit. The loop A (Figure 28) consisted of seven turns of number 16 Brown and Sharp gauge copper wire (1.3 mm. diameter) spaced 10 cm. (4 inches) apart in a solenoidal coil, 30 meters (91.5 feet) long, 7 meters (21.4 feet) high and with the lower conductors of the loop about 4 meters (12.2 feet) above the ground. The open horizontal antenna A_1-A_2 consisted of a single wire 50 meters (153 feet) long, with a variable condenser C_1 inserted at its center. The coils L_1 and L_3 were of 20 millihenrys each, and the resistance R was 1000 ohms. Preliminary tests having indicated that the open antenna received materially less energy than the loop, the resistance R was removed from the open circuit, and placed in the loop. The secondary S was then coupled with L_1 instead of L_3, as shown in Figure 28, and the following readings were taken:

Time P.M.	Station	Loop alone		Loop+Open	
		Signal	Static	Signal	Static
2.45	Nauen	150	300	50	15
2.58	"	300	600	60	5
3.14	Lyons	100	500	30	8
3.28	Nauen	800	3,000	200	10
4.12	Carnarvon	100	500	40	5
4.25	"	100	600	60	8
4.32	Nauen	200	1,000	80	6
4.58	"	100	600	80	10

In conclusion, the writer wishes to acknowledge the valuable assistance and co-operation of Mr. J. A. Proctor, and the aid so freely given by Mr. Alessandro Fabbri, whose helpful enthusiasm and encouragement will long and gratefully be remembered.

SUMMARY: The author considers the possible origins of static and inclines toward a solar origin. He finds that static comes from all directions, with maximum intensity from above and maximum frequency from the horizon. The probable static wave form is a single highly damped pulse.

The author then describes his early work with loop directional receivers, his invention of the so-called "Dieckmann cage," and his early work in static elimination.

He gives the three-dimensional reception characteristics of linear antennas, loops, and various combinations of these.

The work carried on under his direction at the Otter Cliffs, Maine, receiving station is then described in detail. Highly successful loop-and-antenna combinations were used in static elimination, and these are discussed with numerical results.

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15. **Pickard**, "Elec. Rev.," June 15, 1907.
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19. E. **Bellini** and A. **Tosi**, "Elec. Eng.," 2, pages 771-775, November 14, 1907.
20. **Pickard**, U. S. Patent Number 956,165, page 5, lines 37-40.
21. **Pickard**, U. S. Patent Number 876,996, filed June 10, 1907.

DISCUSSION

C. L. Farrand: Mr. Pickard states that a system comprising two loop antennas of Figure 8, United States Patent Specification, Number 956,165, will accomplish the elimination of static providing the currents of the two antennas are properly phased and opposed at the receiver. The reception curve of this system is of the form of a rotated lemniscate or figure of eight, with the points of minimum reception located in a vertical plane perpendicular to the direction of reception, assuming, as is usually the case, that the two antennas are on the line of communication. The system which Mr. Pickard, as he has shown us by the tabulated results of Figure 21, has used so successfully, comprising the earthed loop antenna, has a reception curve approximately of the form of a rotated cardioid, with the points of minimum reception located on the line of communication, in direction opposite to the distant station, and perpendicular to the plane of minimum reception of the two-loop arrangement. It is not apparent, therefore, by Mr. Pickard's explanation how these two systems, which he claims are embodied in his patents, accomplish the elimination of static by "directional reception."

It is to be assumed from Mr. Pickard's remarks that the two-loop arrangement of Figure 8 would function in a locality where the earthed loop of Figure 21 would not.

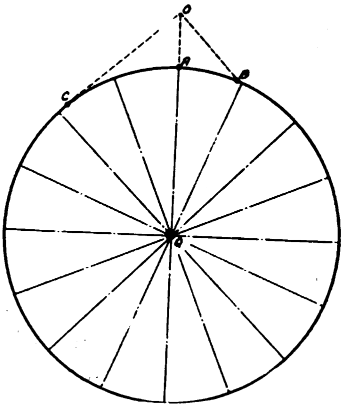
G. W. Pickard: That is approximately correct.

C. L. Farrand: A theory advanced to explain the effect of apparent vertical propagation of static is that the point of origin is at a great height, and that the horizontal components of the static so propagated are neutralized in propagation. Under this hypothesis assume circle $A\,B\,C$ to represent the earth's surface. Static originating at point Q, a relatively great height above the earth, would affect a station located at B as if propagated vertically, but the station located at A would be affected by a static signal of horizontal component $A\,B$ and vertical component $O\,B$ comprising the true direction $O\,A$. The station located at C would be affected by a static signal of purely horizontal direction, Thus static could be eliminated at station B, but stations A and B would be unfortunate in their locations.

A more reasonable hypothesis is that the point of origin of static is approximately at the center of the earth Q, and the radial propagation of static to the surface of the earth affects all stations in substantially the same manner. Thus not only would station *B* eliminate static but stations *A* and *C* would not be slighted.

The determination of direction, vertically upward or vertically downward, may be made by an arrangement similar to that shown by Mr. Pickard's Figure 28. Such tests were under way during the past summer and it is regretted that the shortness of the static season did not permit of their completion.

The hypothesis of vertical propagation has been objected to on the ground that a vertical antenna would accordingly eliminate static and that it is a well-known fact that the vertical antenna receives static very strongly. I might point out that very few antennas have been erected wherein the conductors were vertical or even approximated the vertical. The so-called vertical antenna generally has a considerable horizontal component, or has a flat top of its "L," "T" or umbrella, which horizontal component is responsible for its being affected by static. I have yet to see it proven that a simple vertical earthed antenna will not accomplish the elimination of static. The failure of the vertical earthed antenna may be attributed to the generation of static currents in the conducting earth, the earth in this case representing conductors perpendicular to the direction of propagation and in the path of vertically propagated static. Such is not the case, however, of the vertical unearthed symmetrical antenna, providing that the antenna is elevated sufficiently so that the effect of the earth is negligible. It is hoped that when more complete experimental corroboration can be obtained in various localities and under the many different conditions of service, that the subject matter suggested by the views set forth above may be presented more completely.



Harold H. Beverage: During the past two years, I have observed the static conditions to a considerable extent with a uni-directional receiver known as the "Barrage Receiver."*

When the barrage receiver was first set up at New Brunswick, New Jersey, we found that the signal-stray ratio was most favorable with the adjustments which balanced south-west signals. Furthermore, the signal-stray ratio with uni-directional reception was much better in comparison with the signal-stray ratio on a bi-directional loop than would be expected from a comparison of the horizontal plane intensity characteristics of the two receivers. If we consider that the horizontal plane intensity characteristics of the uni-directional receiver is a cardioid, we would expect the signal-stray ratio to improve in proportion to the reduction in area of the characteristic if the static were equally distributed in all directions.

The improvement to be expected on this assumption would be about 25 per cent. for uni-directional reception as compared with a bi-directional loop receiver. However, we found the improvement to average much more than 25 per cent, often being as much as 300 per cent and occasionally reaching 500 per cent or even more in the afternoon when the static was strong.

A barrage receiver was afterward installed at Otter Cliffs and the same phenomenon was noted there. Recent tests on Long Island have led to the same result. These observations apparently confirm Mr. Pickard's observation that the static is directional on the North Altantic coast with a maximum on the south-western horizon.

At Otter Cliffs, I have observed that, in general, the static in the morning is light and intermittent and does not usually show very great directivity. About 2:00 P.M., the static usually begins to increase rapidly in intensity and continuity, and at the same time begins to show directivity with a maximum in a general south-westerly direction. As Mr. Pickard mentioned, the European stations are roughly north-east from Otter Cliffs and may, therefore, be received with maximum intensity, at the same time preserving a balance on the south-west static. During the summer months, reception would often be impossible at Otter Cliffs with a bi-directional receiver due to strong static, but with uni-directional reception, interruptions are very rare.

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*Described in Mr. Alexanderson's paper, "Simultaneous Sending and Receiving," in August, 1919, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS.

Mr. Pickard's uni-directional receivers at Otter Cliffs played a very large part in maintaining uninterrupted reception from Europe during the war, and he should be congratulated for his splendid contribution to radio communication.

Schenectady, New York,
November 15, 1919.

Greenleaf W. Pickard: Dr. Austin's Figure 9 indicates that he has been misinformed as to Otter Cliffs, as I was at this station during the latter half of August, 1918, and know from my own observation that such a circuit was not in use over this period.[1] Instead, the actual circuit then in use is shown in my recent paper,[2] this circuit having no such open-ended loop as is shown in Figure 9. I am also disappointed to find in this paper no actual comparison of all the various circuits under identical signal and static conditions, that is, side by side. Of course, I sympathize with Dr. Austin in the matter of the cumbersome Lakewood circuit, but I regret that in this instance Mahomet did not go to the mountain.

Dr. Austin agrees with me that a large portion of the static received on the north Atlantic coast comes from the southwest, and this fact suggests a possible connection between thunderstorms and static. A very popular theory of static causation is that its origin is in tropical, or at least southern lightning discharges, the waves resulting from these discharges being reflected back and forth between the earth and the Heaviside layer until they reach the receiving station. It must at once be admitted that this theory has a high degree of plausibility because of the circumstantial evidence in its favor. The seasonal variation in the number of thunderstorms in the United States, according to data collected by our Weather Bureau, may be fairly represented by the curve shown in Figure 1.

This curve shows a maximum in July, and a minimum in the winter months, and is certainly suggestive of static. Even when quantitatively considered, the resemblance is striking, as the ratio of maximum-to-minimum is about 300-to-1, which is also approximately true of static. There are occasional winter days with severe static, and it would be interesting to note these days, and compare them with Weather Bureau records of our southern states for the same days.

As thunderstorms are caused by upward convectional cur-

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[1] "Static Elimination by Directional Reception." Presented before The Institute of Radio Engineers, New York, November 5, 1919.

[2] Previous citation, Figure 25.

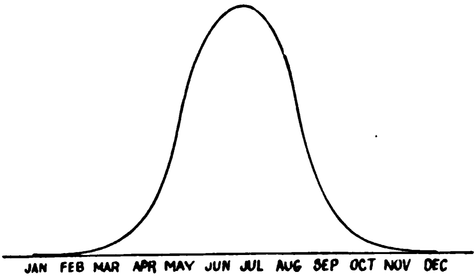

FIGURE 1—Seasonal Variation of Thunderstorms

rents in our atmosphere, they naturally exhibit a diurnal variation in frequency, reaching a maximum shortly after the lower air layers are most heated. Over land areas this will normally be in the afternoon, and we should expect the greatest number of thunderstorms to occur shortly after the greatest heat of the day. In Figure 2 is shown the diurnal variation of frequency of thunderstorms.

From midnight until about eight in the morning, the number of storms falls to a minimum. From then until about four in the afternoon, their frequency increases rapidly to a maximum, later falling off rapidly during the late afternoon and evening. This curve, also, is strongly suggestive of normal summer static,

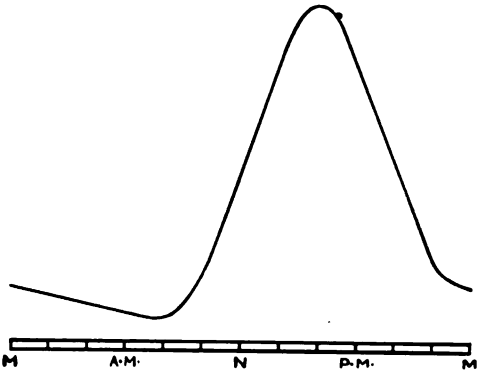

FIGURE 2—Diurnal Variation of Thunderstorms

altho I think the exceptions to the resemblance are more numerous here than with the seasonal curve of Figure 1. So far as trans-atlantic reception is concerned, the late afternoon decline in thunderstorms, if paralleled by static, would be more or less masked by the signal "fading" period. Still, the cumulative effect of these two figures is rather striking, and I do not think the average radio engineer or operator would have questioned these curves, if I had entitled them "Mean Seasonal and Diurnal Variation of Static."

I believe Mr. Marconi was the first to observe that static came from a definite direction. In 1906 he said:[3]

> When using horizontal receiving wires arranged as described in this note, I have often noticed that the natural electrical perturbations of the atmosphere or stray electric waves, which are generally prevalent during the summer, appear to proceed from certain definite directions which vary from time to time. Thus, on certain days, the receiving instruments when connected to wires which are oriented in such a way as to possess a maximum receptivity for electric waves coming from the south, will give strong indications of the presence of these natural electric waves, whilst on differently oriented wires the effects are at the same time weaker or imperceptible. On other days these natural electric waves may apparently come from other directions.
>
> It would be exceedingly interesting to investigate whether there exists any relation between the direction of origin of these waves and the known bearing or direction of distant terrestrial or celestial storms from whence these stray electric waves most probably originate. A considerable number of observations would be necessary to determine whether there exists any relation between the bearing of storm centers and the direction of origin of these natural electric waves. I propose to carry out some further investigations on the subject.

For the past thirteen years I have made directional determinations of static wave-fronts as received at points on the north Atlantic coast, and in general these measurements have indicated a southwestern origin, or, in the alternative, a corresponding plane of polarization for the static waves. During the discussion this evening, Mr. H. E. Campbell mentioned the interesting fact that directional determinations at Goat Island on the Pacific coast, gave a southeastern bearing for static. We are now in possession of cross-bearings, which apparently locate the static source somewhere in the vicinity of the Gulf of Mexico.

Figure 3 is a map of the United States, showing the distribution of thunderstorms during a ten-year period.

[3] "On Methods whereby the Radiation of Electric Waves may be mainly Confined to Certain Directions, and whereby the Receptivity of a Receiver may be Restricted to Electric Waves Emanating from Certain Directions." Read before the Royal Society, March 22, 1906, and published in the "Proceedings," volume LXXVII, -A, pages 413-421. 1906.

There are here two maxima, one over the western coast of Florida, and the other in New Mexico. These two maxima lie southwest from our north Atlantic coast, and southeast from the Pacific, agreeably according to our static bearings.

Having now established at least a *prima facie* case for the connection between static and thunderstorms, we are warranted in examining the matter in greater detail. In the first place, my three previous figures in this discussion appear to rule out of consideration all thunderstorms outside of the United States. If, for example, we consider thunderstorms in Central and South America, it must be noted that as we travel toward the equator, the seasonal variation flattens out, and as we go south of the equator the curve reverses, for the simple reason that South America's summer is our winter, so that these storms cannot be held responsible for our static. In a similar manner, the diurnal variation, and also the cross-bearings I have mentioned above, appear to exclude European or Asiatic storms, because of the difference in longitude or time, as well as bearing. A glance at Figure 3 will show that our worst thunderstorm areas are spread over some two or three hours of longitude, and this should have the effect, so far as static is concerned, of broadening the diurnal maximum shown in Figure 2. Further, as these

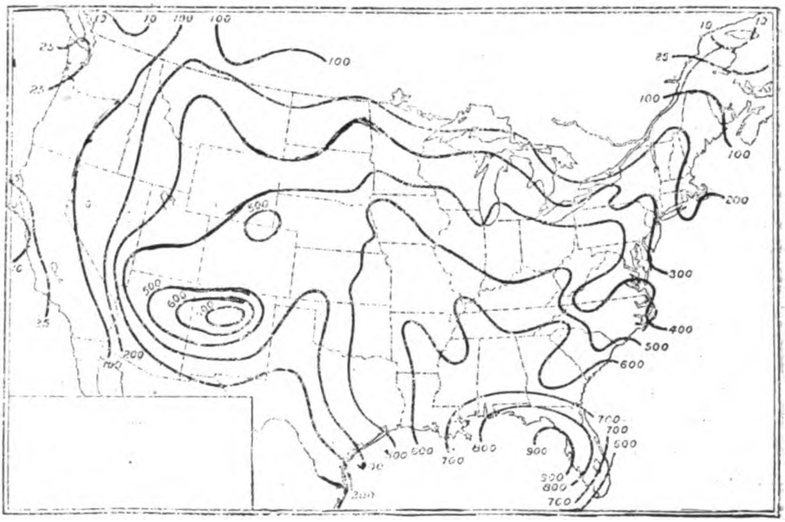

FIGURE 3—Total Number of Thunderstorms During the Ten-year Period, 1904-1913 (Alexander)

thunderstorm areas lie to our west, the effect should be to increase the static during the later afternoon and early evening, as compared with the curve of Figure 2.

Now, during a summer day with severe static, there may average some ten or more disturbances per second, some strong, and others weak, but all going to make up that disagreeable noise complex we call "grinders." Some idea of the number of these disturbances, as well as their varying amplitudes, may be obtained from Dr. Austin's Figure 10, on that portion of the record before my neutralizer system was connected This means from several hundred thousand to perhaps a million disturbances a day, and if each such disturbance is assumed to be due to a lightning discharge somewhere in the United States, the absurdity of the lightning theory is apparent, as a day with, say, half a million lightning strokes would be unique in the annals of meteorology.

Altho, for the reasons set forth above, I do not think that any considerable portion of static is directly due to distant lightning discharges, I am of the opinion that a thunderstorm may determine high level discharges, simply by reason of the disturbance created by its electrostatic field. It is not generally recognized that a thundercloud is not only a very highly charged body, but has a very appreciable electrostatic field many kilometers above itself, nor that thundercloud potentials, and the normal or fair weather potentials, are of entirely different orders of magnitude. My 1906 measurements[4] gave as an approximate thundercloud potential the rather large seeming value of two hundred million volts. C. T. R. Wilson's measurements[5] made by my method some ten years later, give even higher values.

For simplicity, I have first taken a spherical thundercloud, one kilometer (0.62 mile) in diameter and two kilometers (1.24 mile) above the surface of the earth, and charged to a potential of two hundred million volts.

In Figure 4 I have drawn the equipotentials around such a cloud, on the assumption that the conductivity of the atmosphere is negligible. It will be noted that the equipotential of one million volts against ground passes some twenty-odd kilometers (13 miles) above the cloud. Of course no actual thundercloud is spherical, and probably most of them are much larger, but I have used this spherical cloud by way of introduction, and for the reason that it is much easier to compute the equipotentials than with irregular flat clouds. Coming now to something more nearly approaching actuality, Figure 5 shows a thundercloud in the form of a disc one kilometer (0.62 mile) thick, six kilometers (4 miles) in diameter, and two kilometers (1.24 mile) above the ground.

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[4] "The Potential of Lightning," "Electrical World," September 8, 1906, page 491.

[5] "Some Determinations of the Sign and Magnitude of Electric Discharges in Lightning-flashes," "Royal Society Proceedings," 92, pages 555-574, September 1, 1916.

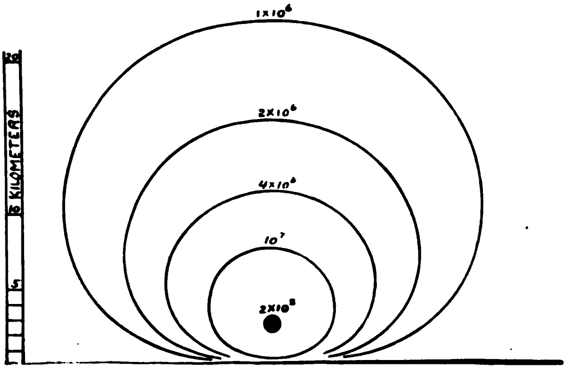

Figure 4—Equipotentials Around Thundercloud

With the charge on the cloud as before, the equipotential of one million volts passes over thirty kilometers (19 miles) above the cloud Actually, severe thunderstorms have cloud areas with mean diameters of over sixty kilometers (38 miles), and it is obvious that the electrostatic field over so extended a cloud area would be even more up-reaching. In fact, for very great heights above a cloud, that is, for distances which are large as compared with the radius and height of the cloud, the potential gradient, where R is the radius of the cloud in centimeters, V the potential of the cloud in volts, and X the distance above the cloud in centimeters, is given by the following simple formula:

$$R^2V/2X^3 \text{ volts per cm.}$$

Of course, it is futile to attempt to calculate these fields with any precision, as we have only an appromimate idea of the potential of the cloud. Further, atmospheric conductivity at the higher levels will modify, that is, flatten out, the upper portion of the field.

The normal or fair weather potential gradients in our atmosphere are quite low as compared with those around thunderclouds, and already have been explored up to about nine kilometers (5.6 miles).[6] A normal potential gradient near the ground is about one volt per centimeter, which decreases logarithmically to about one-thirtieth of a volt per centimeter at a level of nine kilometers (5.6 miles). The atmospheric conductivity,

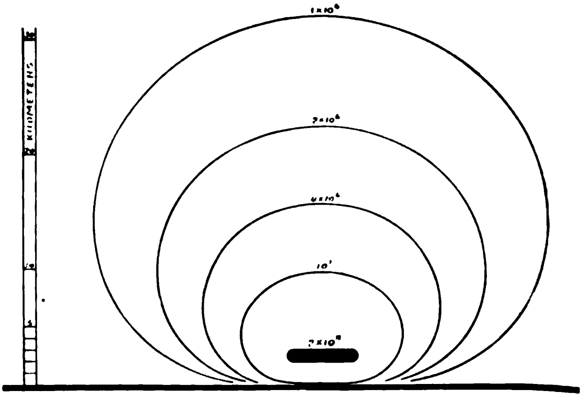

Figure 5—Equipotentials Around Thundercloud

which is about 10^{-4} electrostatic units at the ground level, increases reciprocally with the decrease in potential gradient, at least up to nine kilometers (5.6 miles), and it may fairly be assumed that the entire potential acting across our atmosphere is only of the order of two or three hundred thousand volts. This potential is of course the one responsible for the normal downward electrical currents in our atmosphere, which under normal fair weather conditions are sensibly constant over large areas. It will be at once apparent that a thundercloud, with its enormous charge and field, will profoundly modify these normal conditions, and may bring about a state of affairs somewhat as in Figure 6.

At the left and right of this figure (which is not to sca'e) are the normal equipotential lines, shown dotted, with the normal or fair weather downward currents indicated by the vertical full lines at *A* and *C*. Over the thundercloud *T*, the equipotentials are crowded together, greatly increasing the potential gradient over the cloud, and as a result a much heavier downward current exists at *B*. This depletes the high level charge, at an altitude perhaps fifty or a hundred kilometers (30 to 60 miles) over the cloud, lowering its potential with respect to the normal charge as above *A* and *C*, with the final result that equalizing lateral disharges, as at *D* and *E*, flow in to restore balance. It is my theory that these equalizing discharges are the origin of static, or at least that portion of the total static which is caused by the thundercloud. Of course, for a thundercloud at any considerable distance from the receiving station, the apparent bearing of the static will be that of the cloud itself, but for stations in the south, as at Florida and New Mexico, the bearing will be indeterminate in azimuth, and a considerable portion would apparently come from overhead.

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[6] Wigand and Everling, "Deutsch. Phys. Gesell., Verh.," 16, 5, pages 232-244, March 15, 1914.

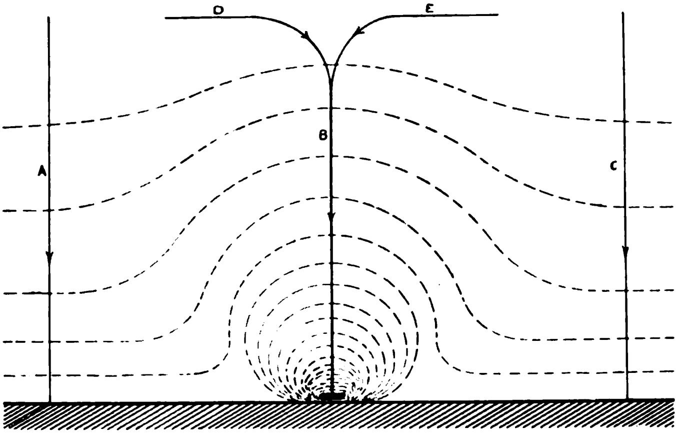

FIGURE 6—Equipotentials and Currents Above Thundercloud

In considering this theory, it may be well to bear in mind that a thunderstorm may be electrically a thunderstorm, even if there is neither cloud, rain nor lightning. Such "electric" storms, accompanied by dry and dusty winds, are observed in fine weather on our western plains, with every evidence of high electrification. Wire fences and other insulated metal objects give shocks when touched, and at night a corona or brush discharge—a St. Elmo's fire—appears at times on pointed objects. It is possible that during the entire thunderstorm season there is a practically constant high potential layer, a few kilometers (a mile or more) above the earth, distributed more or less uniformly over our southern states, and this may determine high level static. On the other hand, it is a rare summer day that does not see a thunderstorm somewhere on our Gulf coast, and one such storm might easily, according to my theory, account for that day's southwestern static.

Louis W. Austin (by letter): I am sorry if I was misinformed in regard to which of the circuits was used at Otter Cliffs during the latter half of August, 1918. I understood that the interwound loop was introduced late in July. The error is of less importance, however, since we know now that the two circuits in question as well as the simple antenna and loop combination all act in the same way, the main differences being in the convenience of adjustment.

It would of course have been more satisfactory to have had all the four systems compared under identical conditions, but the circumstances of war made many things inadvisable which would otherwise naturally have been done. As is pointed out in the paper, the conditions at Belmar, Lakewood, and Washington were considered to be very fairly comparable. Mr. Pickard's theory of the origin of static seems to be plausible tho it is perhaps difficult to understand how electrical disturbances can be produced without visible manifestations, of such magnitude that their waves are more powerful than those of the highest power stations, while a brilliant lightning flash hardly produces a disturbance twenty miles (32 km.) away.

H. H. Beverage: Dr. Austin's classification of the types of static appears to me to be very comprehensive. My experience has been that the first type or grinder type occurs most frequently in the afternoon during the summer months, and is easily balanced for European reception as it almost always comes from a southwesterly direction. I have experimented with very directional antennas, and find that an antenna directive towards the northeast receives very little grinder static, while an antenna directive towards the southwest receives very strong grinder static, often ten to forty times as strong as on a northeast directive antenna of the same dimensions. The second or hiss type static seldom interferes with reception because of its slow variation, which does not produce a disturbance in the radio circuits. The third or click type static does not interfere seriously because of its intermittent character. The click type can usually be traced to direct lightning discharges somewhere in the vicinity of the receiving station.

The fourth type, which sounds something like grinders, but "more crashing in character," is the most serious static remaining after balancing the grinders. The fourth type occurs occasionally in winter, and is particularly bad during the spring months. It shows practically no directivity and cannot be balanced to an appreciable extent by directive receivers, only in so far as the area of the intensity characteristic can be narrowed down towards the ideal form, the search light beam.

At times, there is some evidence that the static of the fourth type is more or less local like the click type static. On Long Island, for instance, during periods when the static of the fourth class was very strong, I have noticed that Belmar would ask for a repeat on a word which was received comparatively free from static on Long Island, while on the other hand, a word almost completely smothered with static on Long Island would be received easily at Belmar. At other times, the same word would be smothered with static at both places simultaneously. It is possible, of course, that static of the third and fourth types existed simultaneously and it was the third or click type that produced the local effects. It is difficult to distinguish between the last two types, as they both produce the same electrical effect except that the fourth type is much more continuous and frequent.

Another experience which may or may not have a bearing on the origin of static was observed in the spring of 1919 on the United States Steamship "George Washington." Radiophone tests were made two or three times daily between New Brunswick and the "George Washington." Two trips were made over and back, and in each case, reception on the "George Washington" was extremely poor in the vicinity of the western edge of the Gulf Stream because of weak signals and very strong static. The New Brunswick radiophone was usually received very well up to three or four hundred miles (500 to 650 km.) and was then practically lost for a distance of about 200 miles (300 km.) when the ship was near the western edge of the Gulf Stream. Beyond this point, New Brunswick radiophone was received again consistently, up to about a thousand miles (1,600 km.), and intermittently, clear to Brest harbor, depending upon atmospheric conditions. In at least one instance, when the static was very bad and the ship was in the Gulf Stream, Otter Cliffs reported no static. European signals were always received much better just outside of New York harbor than in the vicinity of the Gulf Stream. No conclusions can be drawn from so few observations, but this experience may have some bearing on the local nature of some types of static.

C. A. Hoxie: The statement in Dr. Austin's paper that the greater portion of the static disturbances come from the southwest brings to my mind several records I obtained by means of my photographic recorder during the latter part of last August.

At that time, I was not particularly interested as to what direction the major part of these disturbances came from, the main object being to get a record that would show definitely what difference, if any, would be caused by turning the loop, other things being kept the same.

In order to do this two similar galvanometers were installed in one recording instrument and each placed in a position so as to record separately on the same piece of tape as it moved slowly thru the machine, the point of exposure being exactly in line across the tape so it could be readily seen whether or not the recorded impulses were simultaneous. The tape traveled at about 3 inches (7.5 cm.) per second.

Two separate loops, each 15 feet (4.6 m.) square having 25 turns of wire spaced two inches (5.08 cm.) apart were used. The loops were mounted on platforms separated by about 40 feet (12.2 m.) and equipped with wheels so that they could be quickly and easily revolved horizontally to any desired position.

The receiving sets were located midway between the loops, each loop being connected with a receiving set and one of the galvanometers, each receiving set being carefully tuned to the same wave length, approximately 15,000 meters, and adjusted so that the amplification of each and the sensitivity of the galvanometers were practically the same. Three stages of amplification were used, one stage of the resistance type and a two-stage "SE-1,000" Navy type amplifier. Special care was taken to guard against any interference due to tube noises, and so on.

Under these conditions, several records were made and interesting results obtained.

As might be expected when both loops were in the same position, both records were the same. Examination of records of 1, 2, 3, and 4 will show this. Records 5, 6, 7, 8, and 9 show that impulses that affect one loop do not necessarily affect the other if placed at a different angle.

An inspection of 6, 7, and 8 brings out two interesting points. 1st:—Considerably more static was recorded from the loop located extending NE-SW than from the loop placed E and W. 2nd:—Quite strong impulses were recorded on each of the two loops that were not recorded on the other.

An inspection of record 9 shows that much more static was recorded on the loop placed NE-SW than the one placed NW-SE.

In order more definitely to determine whether the greater portion of static came from some general direction and also to get some idea of the relative strength, it was decided to use only one loop. This was done to eliminate the chance of the amplification of one set being larger, or the sensitivity of one galvanometer being more than the other.

The procedure was as follows. The loop was placed, say, in a north and south position and a half minute record made (approximately 5 feet (1.5 m.) of tape); it was then quickly moved to NE-SW position and another half minute record made, then to E and W, and lastly to SE-NW.

Several of these tests were made on different days and showed very conclusively that the static effect was greatest when the loop was in a NE-SW position and the least when in a NW-SE position. Also that the static effect was greater when the loop was turned east-and-west than when it was in a north-and-south position.

Special care was taken during each test that no change was made in tuning, amplification, or the galvanometer adjustment.

The vibrating reed of the recorder is not aperiodic, but is tuned to 2,000 cycles. It was adjusted so as to reach full amplitude with 3 cycles, requiring the same time to come to rest, so that an impulse of oscillation lasting 0.0015 second would give full amplitude coming to rest 0.0015 second later, the total elapsed time being only 0.003 second, and representing 0.01 inch (0.025 cm.) on the tape.

Keeping in mind the above statement that not more than 0.01 inch (0.025 cm.) in the tape is required for the building up and damping after the cessation of an impulse, it is evident that if any more space is taken up it must be due to the impulse persisting longer than the 0.0015 second necessary for reaching full amplitude; or if less space than 0.01 inch (0.025 cm.) is taken up on the tape, the duration of the impulse was evidently less than 0.0015 second.

Records 10, 11, 12, and 13 are the results of one test and show very nicely the relative values of static with the loop in each of the four positions.

Records 14, 15, 16, and 17 show results of one other of the numerous tests made.

Only a few records of the many made are shown here, but I believe they can be considered typical of the whole.

A. S. Blatterman: Some observations on the direction of static have been made by the Signal Corps. In 1918, observations made at the Radio Laboratories, Camp Alfred Vail, New Jersey, showed that there were certain periods when the direction of the disturbances was ill defined or even impossible to determine altogether. Most of the time, however, it was easy to swing a loop and find a marked direction, and when this was possible the direction indicated was always southwest-northeast. This agrees with statements made in the paper, and with observations of Mr. Hoxie and others who have just discussed the paper.

During the summer of 1919, I made some similar observations on the directivity of static, while in the Army, at El Paso, Texas. These observations were made on a five-foot (1.53 m.) square loop, with a non-oscillating receiver, on wave lengths from 400 to 1,000 meters. During a period of ten days, during which time the weather at El Paso was continuously clear, it was impossible to find any directional effect at all.

Since El Paso is right on the border between Texas and New Mexico, and therefore quite close to the southwest thunderstorm center described by Mr. Pickard, the above observations give support to his theory. A receiver located at the origin of the static would not be expected to show direction on these disturbances.

Bowden Washington: I am curious to know the fundamental reasons for attempting to balance ground wires, water wires, and so on, against loops. It is, of course, realized, that the signal-static ratio in these two devices will probably be different and there is the possibility of balancing out the static and having some signal left, but I cannot but feel that it is starting under a difficulty.

If the static is to be balanced out, the phase should be the same, or rather it will be preferable to have it the same so as to eliminate phase-changing devices. The damping of the two circuits should be similar so that the envelopes will be similar. It is, of course, realized that when trying to balance two dissimilar energy receiving systems, it is possible by varying the constants of the circuits and by changing some phase-shifting device finally to accomplish this result; but all systems which consist of an effort to balance two dissimilar energy collectors appear to be starting under a handicap. It will certainly seem less difficult to obtain a balance between two similar devices in which the damping, receptive ability, and so on, are equal in the first place and in which the phase of the desired wave is different.

I should like to make a few remarks on the method of comparison between these various devices. I realize that to a certain extent the Navy had to use such locations and materials as were available and that research work had to be made secondary to the handling of traffic. However, I cannot but believe that comparisons made between apparatus situated in widely different localities, over somewhat different distances of transmission and with different personnel are almost valueless. It can almost be said without exaggeration that the personnel factor alone must constitute the biggest variable. It seems to me that the one true test is when the various devices to be tested are in the same locality, at the same time, receiving from the same transmitter, and, if recorders cannot be used, that the personnel should be interchanged so frequently as to reduce this variable. The reason I am bringing this point out is that it has been my experience with the Navy Department that it frequently commits this error, that in making a test to determine the ratio between two dependent variables, they take very little account of what is going on with a dozen or more dependent variables all of which are inter-related.

For instance, the writer witnessed a "range test" of transmitters in which the transmitters were operated from two different craft on different days with different receiving antennas and different operators—in one case a mineral rectifier was used for reception, while in the other a three-electrode valve. A direct comparison was made between the ranges obtained during these two tests.

I cannot but feel that in an art as new as radio engineering, in which we do not know the relations between a great many variable factors, it is most important to make every effort to keep constant, or I might say, keep equal, every variable factor, except the one directly on test.

S. M. Kintner (by letter): Dr. Austin's extremely interesting paper gives facts that should serve to convince even the most skeptical that long distance radio communication has been demonstrated. How many of the ocean cables can show continuous operation for practically 99 per cent of the time? Yet that is the result indicated for radio by Dr. Austin as having been accomplished. None of those familiar with radio operations doubt for one instant the ability of radio to render such service; yet few, if any, of the cable interests are willing to make any admissions regarding the capabilities of radio as a competing means. This is perhaps not strange;—it is but human nature. None are so blind as those who have eyes but will not see. When we look back on other great developments that have paralleled existing means of accomplishing substantially the same results, we are impressed with the similarity of the reception accorded them by the existing means and that given the radio art by the cable interests. It is also a rather significant fact that these new arts were not developed by those already in the business, but by entirely new interests. For instance, the early electric light companies were not promoted by the gas companies. The electric street cars were not developed by the old horse car operators. The automobile taxi-cab companies were not started by horse-drawn cab owners. The telephone even was not considered of much commercial utility by the existing telegraph companies, when it was first introduced to the public.

It is not strange then that human nature should again display the same characteristics and cause those now operating cables to close their eyes to the possibilities and capabilities of the greatest competitor they have ever had and refuse to see the radio art in its true importance. It is to such people that Dr. Austin's paper should be most instructive.

In the second paragraph of the paper and in a footnote referring thereto, reference is made to the oscillating audion. It is rather unfortunate that this should not have been described as the heterodyne receiver. The oscillating audion *per se* is of value in accomplishing the result described only when operating in the heterodyne arrangement of which it is but one form.

The general data supplied is most opportune and of great interest to the radio art.

Frederick K. Vreeland: These questions concerning the origin and direction of propagation of strays that have been raised by Mr. Pickard and other members are very interesting, and I say by all means let us learn all we can about them. But in the last analysis, this knowledge will not give us the final answer to the problem of stray elimination.

The practical question that interests the radio engineer is not the origin of the strays, but what do they do when they strike our collecting circuit, and how can we prevent them from doing it.

Dr. Austin has mentioned the four generally-accepted classes of strays, describing them according to the sounds which they produce in the receiver. We can all recognize them by these sounds, but if we wish to handle them intelligently we must know, in addition, what kind of currents they produce in the collecting circuit. It may be well, therefore, to review briefly our knowledge of this subject.

Taking the simplest case first, the frying and sizzling disturbances accompanying rain or snow are static, properly so called, and are due to potential differences existing in the neighborhood of the receiver. If we put a direct current microammeter in the antenna circuit, it will show a deflection, sometimes positive, sometimes negative, and sometimes reversing. This current may sometimes reach a surprisingly large value, but it produces little effect on the detector since the detector responds only to the changes in the current. Since these changes are relatively slight and not very rapid, the disturbing effect on the receiver is small.

Strays caused by nearby lightning discharges, on the other hand, induce enormous surges in the antenna. When the lightning flash occurs the induced charge on the antenna is suddenly released, producing shock oscillations of great intensity. These oscillations may jump the safety gap, but they produce comparatively little interruption of the reception since they are of short duration and far between.

The class of "grinders" are as a rule noisy, but really not as bad as they sound. They consist in rapidly recurring impulses of relatively small amplitude but following each other in quick succession. While they are so noisy as to produce serious interference with the reception, they are not so difficult to eliminate as some other strays. It is just as easy to eliminate one hundred stray impulses per second as it is to eliminate one, tho the hundred make a far louder noise. This is why some stray-reducing systems appear to a superficial observer to produce an extraordinary effect in taking out the noise of these grinding strays while failing utterly to conquer the apparently inconspicuous but much more troublesome clicks and crashes. One does not need to use an apparatus six miles (10 km.) long to conquer these grinders. That can be done in a little box two feet (60 cm.) on the side that stands on the operating table.

The really vital problem of stray elimination, therefore, reduces to the question of handling the so-called clicks and crashes. These are characterized by their very sudden impulsive nature and short duration. They are either aperiodic or very strongly damped, and because of their extreme abruptness and transitory character, they have an extraordinary power of exciting shock oscillations.

The vital problem of stray elimination, therefore, reduces to the prevention of such shock oscillations or making them innocuous in affecting the receiver.

I think it must be obvious to us all that a directional system alone cannot give the final solution of the problem, for however effective it may be in eliminating strays coming from a given direction, there will always remain a residuum coming from other directions, and this is sufficient to cause serious disturbance.

A balancing system, such as those described by Dr. Austin, to be fully effective must balance the shock oscillations in the two collectors, not only in magnitude but also in phase, in frequency, and in decrement.

Obviously if the two effects are not of the same amplitude, they will not balance.

Clearly, also, if they are not in phase, they will not cancel each other.

If they have not the same frequency, there will be certain intervals when they will be out of phase and hence will not balance.

If they have not the same decrement, there will be certain intervals when they will have different amplitudes and so will not balance. To secure perfect coincidence in all these four respects is not easy, and requires more than the simple balanced circuits described in Dr. Austin's Paper.

There are, however, simpler means of doing away with the disturbing effects of these sharp impulsive strays. This is not the time nor the place to explain how it may be done. That is another story.

Louis W. Austin (by letter): Of course Mr. Vreeland is right in remarking that balanced systems cannot afford a complete solution of the static problem since frequently local non-directional static is capable of producing considerable trouble. Then, too, there are cases when directional static comes from more than one center at once, as recently observed in Porto Rico; but in general, at least along the Atlantic Coast of the Northern United States, the directive systems are very effective especially in reception from Europe.

I cannot agree with him in regard to static being always a pure shock phenomenon, as recent experiments which will be published later indicate that often static pulses do not appear simultaneously on the different wave lengths, showing that the static has definite wave lengths of its own forming probably in some cases a continuous spectrum.

Of course we shall all be delighted if Mr. Vreeland is able to furnish a simple device to be connected to any receiving system which will entirely solve the problem.

QUANTITATIVE EXPERIMENTS WITH COIL ANTENNAS IN RADIO TELEGRAPHY*

By

Louis W. Austin

(United States Naval Radiotelegraphic Laboratory,
Washington, D. C.)

The use of large inductance coils for sending, receiving, and direction determination was first proposed by the late Professor Braun[1], who carried out experiments at Strassburg on signals from the Eiffel Tower, and gave the general theory of a coil used as an antenna.[2]

Received current measurements, such as the laboratory made some years ago for the verification of the theory of transmission between ordinary antennas, have now been made for closed coils. According to the theory, a rectangular vertical coil of N turns, height H, and length L, is equivalent to two vertical antennas of effective height NH at a distance apart L, with their respective currents in opposite phase. Taking into account the phase difference due to the difference in path, either in sending to a point p, or in receiving from p, the effect will be the same as that of one antenna of height NH multiplied by the phase difference $2\pi \frac{L}{\lambda} \cos \theta$, where θ is the angle between the plane of the coil and the direction of p.[3]

Now the expression for the received current in a receiving antenna, with an antenna sending sustained waves, is, disregarding absorption:[4]

$$I_r = 120\pi \frac{I_s h_s h_r}{\lambda d R} \tag{1}$$

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*Received by the Editor, July 10, 1919. Presented before The Institute of Radio Engineers, New York, May 5, 1920.

† Dr. Dellinger, of the Bureau of Standards, has published coil formulas in practical units with slightly different constants in a confidential report for the Signal Corps, "Radio Transmission Formulas," July, 1917.

[1] F. Braun, "Jahrbuch d. Drahtlosen Telegraphie," 8, pages 1 and 132, 1914.

[2] Mr. Kolster, of the Bureau of Standards, has developed an excellent direction finder on this principle.

[3] Zenneck, "Wireless Telegraphy," 1915, page 234 and note 307, page 425.

[4] Zenneck, "Wireless Telegraphy," 1915, page 248, and "Bureau of Standards Bulletin," 11, page 70, 1914, Scientific paper 226

where I_s is the sending antenna current, h_s the effective height of the sending antenna, h_r that of the receiving antenna, λ the wave length, d the distance, and R the receiving antenna resistance, expressed in amperes, ohms, and meters. Now if either or both of the antennas is replaced by a coil, we must replace its height by $NH \cdot 2\pi \frac{L}{\lambda} \cos\theta$ where NH is the height of the coil times the number of turns, and we find for a

Coil Sending and Antenna Receiving

$$I_r = 120\pi \cdot \frac{I_s N_s H_s}{\lambda d} \cdot 2\pi \frac{L_s}{\lambda} \cos\theta_s \cdot \frac{h_r}{R}$$

$$= 2369 \cdot \frac{I_s N_s H_s L_s h_r}{\lambda^2 d R} \cdot \cos\theta_s \qquad (2)$$

In the same way for an

Antenna Sending and Coil Receiving

$$I_r = 2369 \cdot \frac{I_s h_s N_r H_r L_r}{\lambda^2 d R} \cos\theta_r \qquad (3)$$

and for

Coil Sending and Coil Receiving

$$I_r = 14880 \cdot \frac{I_s N_s H_s L_s N_r H_r L_r}{\lambda^3 d R} \cdot \cos\theta_s \cos\theta_r. \qquad (4)$$

The effective height h of an ordinary antenna equivalent to any coil NH may be expressed if $\theta = 0$, by

$$h = 2\pi \frac{NHL}{\lambda} = 2\pi \cdot \frac{area \times turns}{\lambda} \qquad (5)$$

The equations show that, other things being equal, if an antenna be used, both for sending and receiving, the received current falls off as the wave length, while if one coil be used, it falls off as the square of the wave length, and with two coils as the cube of the wave length.

The value of the constant in the equation for a sending coil requires some consideration. The value given assumes that as the radiated field grounds itself, it takes the form of a field formed by the coil and its image as in the case of an antenna. This is probably true, at least for a coil the dimensions of which are large compared with its distance from the ground. In the case of reception, this question does not enter in.

Signals from Arlington have been measured at the laboratory on two coils. One was a crossed coil direction finder mounted on the roof of one of the buildings, having 56 closely wound turns, a height of 1.82 meters (5.97 ft.) and a length measured between the planes of the front and back vertical sections of 1.29 meters (4.23 ft.). The second coil, supported from masts, had 7 turns 80 cm. (2.62 ft.) apart and measured 21.6×24.4 meters (71×80 ft.), θ being 42°. The results are shown in Tables 1 and 2.

TABLE 1

Arlington Arc Received at Laboratory on Direction Finder.

λ meters	Received Current amp. obs.		Difference	%Difference
	obs.	calc.		
6,000	$208.0 \cdot 10^{-6}$	$157.5 \cdot 10^{-6}$	50.5	24%
7,500	128.1	100.8	27.3	21%
10,000	67.5	56.7	10.8	16%
$I_s = 100$ amp.	$N = 56$		$L = 1.29$ m.	$\theta = 0°$
$h_s = 71$ m.	$H = 1.82$ m.		$R = 50$ ohms	$d = 7{,}800$ m.

TABLE 2

Arlington Arc Received at Laboratory on Large Coil.

λ meters	Received Current amp.		Difference	%Difference
	obs.	calc.		
4,000	$8.78 \cdot 10^{-3}$	$7.38 \cdot 10^{-3}$	1.40	16%
6,000	3.71	3.28	0.43	12%
7,500	2.26	2.10	0.16	7%
10,000	1.23	1.18	0.05	4%
$I_s = 100$ amp.	$N = 7$		$L = 24.4$ m.	$\theta = 42°$
$h_s = 71$ m.	$H = 21.6$ m.		$R = 50$ ohms	$d = 7{,}800$m.

Table 3 shows the current received on an antenna at Arlington from the large coil at the laboratory excited by a coupled bulb circuit. The received currents in all three cases were measured with shunted detectors and galvanometers,[5] calibrated and tested for proportionality between deflection and current squared in each experiment.

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[5] "Journal Wash. Acad.," 8, page 569, 1918.

TABLE 3

Large Coil at Laboratory Received on Antenna at Arlington·

λ meters	Received Current amp. obs.	Received Current amp. calc.	Difference	% Difference
2,800	$147.8 \cdot 10^{-6}$	$150.8 \cdot 10^{-6}$	-3.0	2%
4,890	50.5	49.5	$+1.0$	2%
$I_s = 1$ amp.	$N = 7$		$L = 24.4$ m.	$\theta = 42°$
$h_r = 71$ m.	$H = 21.6$ m.		$R = 50$ ohms.	$d = 7{,}800$ m.

In each of the tables, for the sake of comparison, the results are reduced to a common value of sending current and a common receiving resistance.

The observed values in Tables 1 and 2 are seen to be uniformly larger than the calculated. This is supposed to be due to an action of the coil as an antenna since an increase in the length of the leads increases this difference. The reason for the increased error at the shorter wave lengths is not yet clear.[6] The agreement of observed and calculated values in the case of the coil sending (Table 3) is all that could be desired. From this, sending from a coil and receiving on an antenna, seems to offer the most accurate method for determining antenna effective height.

Table 4 gives the effective heights of antennas which for sending or receiving are equivalent to coils of various area turns, calculated from Equation 5.

When I wrote this paper, I supposed that Messrs. R. A. Fessenden and G. W. Pickard used only single turn loops in their early experiments, but I have recently been informed that Professor Fessenden proposed multiple turn loops in his German patent number 225226, dated January 14, 1907, while Mr. Pickard demonstrated the use of a direction finder of this type to the Navy in the same year, and published his work in the "Electrical Review," June 15, 1907, and October 3, 1908. That these early experiments did not result in the general use of receiving coils was due entirely to the lack of sensitive receiving apparatus at that time, while their present success is due to audion amplifiers, the introduction of which practically coincided with the appearance of Professor Braun's paper.

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[6] In experiments of this kind incorrect results may be obtained if observations are made too near the natural wave length of the coil, as in this case the current distribution is no longer uniform on account of the effect of the distributed capacity

TABLE 4

Comparison of Coils and Antennas

λ (meters)	Equivalent Effective Antenna Height (meters)				
	10	20	30	40	50
	Coil-Area × Turns (meters)				
200	319	—	—	—	—
500	796	1,592	2,388	—	—
1,000	1,592	3,184	4,776	6,370	—
2,000	3,185	6,370	9,552	12,740	15,920
3,000	4,776	9,552	14,280	19,100	23,880
5,000	7,960	15,920	23,880	31,840	39,800
7,000	11,140	22,280	33,450	44,580	55,760
10,000	15,920	31,840	47,760	63,700	79,600
12,000	19,110	38,220	57,330	76,440	95,520
15,000	23,880	47,760	71,640	95,520	119,400
20,000	31,840	63,780	95,520	127,400	159,200

The observations in this paper have been taken for the most part by W. F. Grimes, Chief Electrician (Radio), assistant in the laboratory.

United States Naval Radio Laboratory,
Washington, D. C.
April, 1919

SUMMARY: After deriving formulas for received current in loop receivers or antennas (transmission being from other loops or antennas), these formulas are compared with a series of observations at various wave lengths. Good agreement is found with theory, and the residual discrepancy is partly explained.

DISCUSSION

J. H. Dellinger (by letter): The difference in the values of the constants of the transmission formulas as given by Austin and by me ("Bureau of Standards Scientific Paper," number 354) require a word of explanation, especially since the equations have been printed side by side in the "Radio Review," without comment.

The Austin antenna-to-antenna formula as given in his later papers, except for the attenuation term, is identical with the purely theoretical formula for the Hertzian oscillator, the ground being supposed to be a perfect conductor so that the antenna and its image form a hypothetical oscillator of a length equal to twice the height of the antenna to its center of capacity. My antenna-to-antenna formula is the theoretical formula for the Hertzian oscillator having its length the actual height of the antenna above the ground. In other words, Austin assumes a perfect image below the antenna and I assume no image. The truth is somewhere between the two.

There is no question that my formulas are the more nearly correct for any case where the radiating structure is away from the ground, as an airplane antenna of any type, or a condenser antenna (as described in my paper) or a coil antenna raised well off the ground.

For an antenna on very good conducting ground or on a ship in salt water there is probably a considerable image effect. In practice, on account of the partial effect of the earth conductivity, the presence of metal towers and so on, the image effect is partly lost, so that the equivalent oscillator is generally less than twice the antenna height. This is taken account of by Austin by defining h_s not as the actual but as the effective height of the antenna, that is, half the length of the oscillator which will produce the same radiated field as the antenna at a distance. It is found in many cases that the effective height is about half the total height and seldom approaches the full value. For the convenience of those using the formulas I have taken h as the full antenna height, since this is the simplest procedure and leads to results in agreement with experiment.

Dr. Austin agrees with me in the foregoing explanations.

George S. Davis (by letter): While it is true that the early experiments referred to by Dr. Austin did not immediately result in the general use of coil aerials, still I do not think that this was due entirely to the lack of sensitive receiving apparatus at that time but rather to the incredulity with which the announcement was received that one could determine by the use of a coil aerial the bearing or direction of the distant transmitter. I was in charge of the radio station at the Brooklyn Navy Yard during 1907 when this announcement was made and my recollection is that we treated it in much the same manner as we did announcements to the effect that someone had entirely overcome or eliminated static.

Of course, audion amplifiers have added materially to the utility of the coil aerial, but the fact remains that coil aerials are useful without audion amplifiers and could have been brought into universal use in 1907 had the announcement been given due credence or had Mr. Pickard at that time possessed the means to push it, or had the necessities later brought about by the war compelled the various governments, including our own, to adopt it for submarine detection and other purposes at that time.

Perhaps one of the most controlling factors contributing to the general use of the coil aerial was the absolute necessity of detecting the location of enemy transmitting stations at the front and on submarines, and to these necessities I think is due the general adoption of the coil aerial as much as to the audion amplifiers.

As distinguished from the necessities of the war, our problems and necessities in 1907 were altogether different and there were other problems remaining unsolved which at the time probably seemed more important to the "powers that be" than the problem of determining direction, altho in the light of what we know now, it is hard to understand why we in the United States Navy did not more vigorously follow up the results demonstrated to (then) Lieutenant Robison in 1907 by Mr. Pickard in the matter of coil aerials.

I also recall that during a visit to Brant Rock, Massachusetts, in 1910, Professor Fessenden was then conducting experiments which I understood were direction-finding experiments from his station at Plymouth, and he at that time urged upon Mr. Musgrave of the United Fruit Company the adoption of an antenna structure in the United Fruit Company's system of stations which would, as he explained, be directional both as to transmission and reception and would thus not only reduce interference but result in having to use less power for communication over a given distance. I do not recall whether the arrangement used by Professor Fessenden was that proposed in his

German patent mentioned by Dr. Austin, but, as this patent was first published in 1910 instead of January 14, 1907, as stated by Dr. Austin, it is possible that the arrangement used was that shown in the Fessenden patent. As I recall it, Fessenden was reported to have secured fairly good directional results in 1910, but here again the so-called direction-finding aerials did not come in general use, due not to any lack of sensitive receivers, but to the lack of demand for such aerials on the part of commercial interests in radio at this time and the apparent lack of necessity in Government service which was then perhaps the largest field for radio apparatus. Even after the Braun 1914 publication, when we had amplifiers, coil aerials did not come into immediate general use, due again to lack of demand, and it was not until 1915-16 when the absolute need for direction-finding stations was demonstrated through war operations that coil or direction-finding aerials were generally introduced. In fact, it seems that the English radio authorities did not give radio direction-finding aerials or apparatus much thought until after the battle with the Germans near the Falkland Islands, when it was generally reported that the Germans had located the English fleet through means of a radio direction-finding apparatus. In any event, it is notable that, after this fight, the adoption of radio direction-finding apparatus was more intensive than ever before.

In conclusion, it seems to me that, altho the audion amplifiers have contributed to the success of the coil aerials as demonstrated by Pickard in 1907, it was the necessities of war which brought them into general use, and pointed the way generally for their commercial adoption.

A DISCUSSION ON "ELECTRICAL OSCILLATIONS IN ANTENNAS AND INDUCTION COILS"* BY JOHN M. MILLER

By
AUGUST HUND
(UNIVERSITY OF CALIFORNIA, BERKELEY, CALIFORNIA)

Dr. Miller's interesting paper[1] on "Electrical Oscillations in Antennas and Inductance Coils" in the June, 1919, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS is undoubtedly a paper of importance to the radio engineer, since there exists quite a confusion with respect to the proper determination of the various antenna constants and the different types for each constant. Altho this subject has been treated successfully in the earliest days of the art of radio engineering, changes have occurred with respect to the design of antennas and their excitation. Long horizontal antennas are now used more often and sometimes excited by means of sustained sinusoidal waves on account of the successful devolopment of suitable ferric and non-ferric radio sources.

Professor Morecroft's investigations[2] on an artificial antenna, operated at commercial frequencies, was therefore a timely contribution, since he demonstrated—in a most scholarly way—the distribution of the net impedance and confirmed the resonance current as well as the proper oscillation constant of an actual antenna by means of an artifical closed circuit.

The purpose of the following discussion is to give formulas for the calculation of the correct effective antenna constants for any coil-loading whatever for the condition of a long horizontal antenna excited by a sinusoidal radio frequency electromotive force if the electromagnetic field is under consideration. The results are compared with the formulas giving only the apparent effective constants. Such a comparison will then lead to a method by which the static constants as well as the correct effective antenna constants can be determined, and by means of which the effective coefficient of the loading inductance can be cal-

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* Received by the Editor, August 1, 1919.

[1] Volume 7, page 308 to page 311.

[2] "Some Experiments With Long Electrical Conductors," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, December, 1917, volume 5, page 389 to page 411.

culated also. The deductions make use of the distributions indicated in the figure and the radio equation for the horizontal aerial.

The formulas for the correct antenna constants for the elecmagnetic point of view are

$$\left.\begin{aligned}
C_{e1}' &= \frac{8\lambda'}{\pi\lambda}\,\frac{\sin^2 90\frac{\lambda}{\lambda'}}{\pi\frac{\lambda}{\lambda'}+\sin 180\left(1-\frac{\lambda}{\lambda'}\right)}\,C_A = A_1 C_A \\
L_{e1}' &= \frac{\lambda'}{2\pi\lambda}\,\frac{\pi\frac{\lambda}{\lambda'}-\sin 180\left(1-\frac{\lambda}{\lambda'}\right)}{\sin^2 90\frac{\lambda}{\lambda'}}\,L_A = B_1 L_A \\
r_e' &= B_1 r_A
\end{aligned}\right\}\quad (1)$$

while for the apparent effective constants they are

$$\left.\begin{aligned}
C_{e2}' &= \frac{2\lambda'}{\pi\lambda}\sin 90\frac{\lambda}{\lambda'}\,C_A = A_2 C_A \\
L_{e2}' &= \frac{2\lambda'}{\pi\lambda}\,\frac{\left(1-\cos 90\frac{\lambda}{\lambda'}\right)}{\sin 90\frac{\lambda}{\lambda'}}\,L_A = B_2 L_A
\end{aligned}\right\}\quad (2)$$

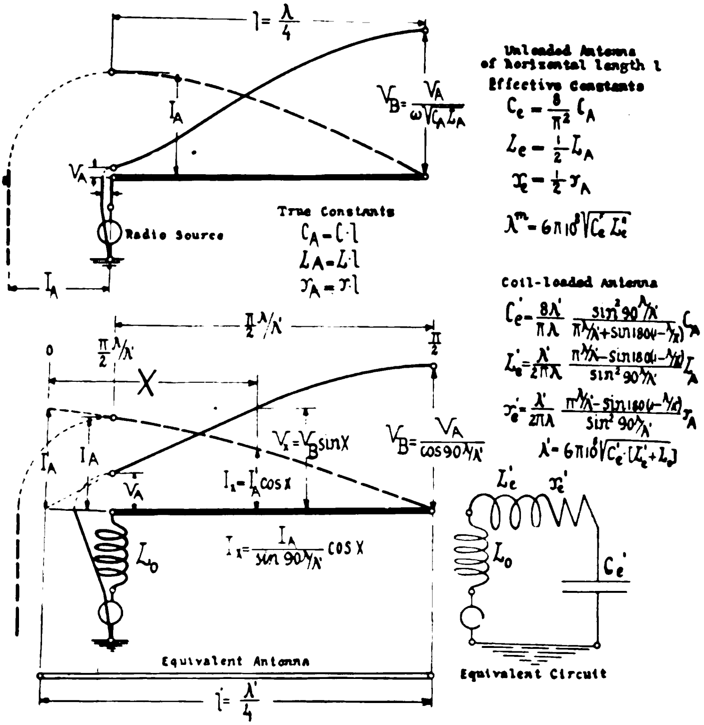
if λ and λ' denote the natural fundamental wave lengths for the unloaded and the coil-loaded antenna system, respectively. The quantities C_A, L_A, and r_A represent the static or true antenna constants for which the potential, as well as the current are, for all points of the horizontal length l, of unchanged amplitude and same phase. Any low frequency method will then give the values

$$C_A,\quad \frac{L_A}{3},\quad \frac{r_A}{3}$$

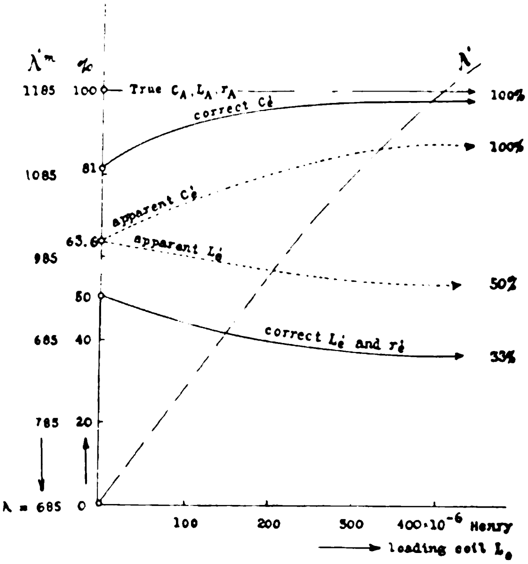
since then the effective antenna potential is practically constant along the aerial and the effective current decreases almost linearly to its zero value at the open end of the aerial.

The correct values, C_{e_1}', L_{e_1}' and r_e', for the effective capacity, inductance and resistance of the antenna will, for an inserted loading inductance L_o, confirm the oscillation constant

$$C_{e_1}'\,[L_{e_1}'+L_o]$$



$l = \frac{\lambda}{4}$
I_A
V_A
$V_B = \frac{V_A}{\omega\sqrt{C_A L_A}}$
Radio Source
True Constants
$C_A = C \cdot l$
$L_A = L \cdot l$
$r_A = r \cdot l$
Unloaded Antenna of horizontal length l
Effective Constants
$C_e = \frac{8}{\pi^2} C_A$
$L_e = \frac{1}{2} L_A$
$r_e = \frac{1}{2} r_A$
$\lambda^m = 6\pi 10^8 \sqrt{C_e L_e}$
Coil-loaded Antenna
$C_e' = \frac{8\lambda'}{\pi\lambda} \frac{\sin^2 90\lambda/\lambda'}{\pi\lambda/\lambda' + \sin 180(1-\lambda/\lambda')} C_A$
$L_e' = \frac{\lambda'}{2\pi\lambda} \frac{\pi\lambda/\lambda' - \sin 180(1-\lambda/\lambda')}{\sin^2 90\lambda/\lambda'} L_A$
$r_e' = \frac{\lambda'}{2\pi\lambda} \frac{\pi\lambda/\lambda' - \sin 180(1-\lambda/\lambda')}{\sin^2 90\lambda/\lambda'} r_A$
$\lambda' = 6\pi 10^8 \sqrt{C_e' \cdot [L_e' + L_o]}$
$\frac{\pi}{2}\lambda/\lambda'$
0
$\frac{\pi}{2}\lambda/\lambda'$
$\frac{\pi}{2}$
X
I_A
I_A
V_A
$V_x = V_B \sin X$
$V_B = \frac{V_A}{\cos 90\lambda/\lambda'}$
$I_x = I_A' \cos X$
$I_x = \frac{I_A}{\sin 90\lambda/\lambda'} \cos X$
L_o
L_e'
r_e'
L_o
C_e'
Equivalent Antenna
Equivalent Circuit
$l' = \frac{\lambda'}{4}$



λ^m
%
λ'
1185
100
True $C_A \cdot L_A \cdot r_A$
100%
correct C_e'
1085
81
100%
apparent C_e'
985
63.6
apparent L_e'
50%
60
885
40
correct L_e' and r_e'
33%
785
20
λ = 685
0
100
200
300
$400 \cdot 10^{-6}$ Henry
loading coil L_o

of the antenna as well as the surge impedance

$$\sqrt{\frac{L_{e_1}'+L_0}{C_{e_1}'}}$$

since the artificial antenna made up of these constants will produce the same resonance current for both damped and undamped sinusoids. That means, the correct effective constants will verify the generalized impedance

$$Z=r_e'+n\left[L_{e_1}'+L_0\right]+\frac{1}{C_{e_1}'}=0$$

for the generalized angular velocity $n=a\pm j\omega$, since the constants are based on the derivations

$$\left.\begin{aligned}
C_{e_1}' &= \frac{\left[\int_{\frac{\pi}{2}\left(1-\frac{\lambda}{\lambda'}\right)}^{\frac{\pi}{2}} C\sin x\,dx\right]^2}{\int_{\frac{\pi}{2}\left(1-\frac{\lambda}{\lambda'}\right)}^{\frac{\pi}{2}} C\sin^2 x\,dx}\\
L_{e_1}' &= \int_{\frac{\pi}{2}\left(1-\frac{\lambda}{\lambda'}\right)}^{\frac{\pi}{2}} L\frac{\cos^2 x}{\sin^2\frac{\pi}{2}\frac{\lambda}{\lambda'}}\,dx\\
r_e' &= \int_{\frac{\pi}{2}\left(1-\frac{\lambda}{\lambda'}\right)}^{\frac{\pi}{2}} r\frac{\cos^2 x}{\sin^2\frac{\pi}{2}\frac{\lambda}{\lambda'}}\,dx
\end{aligned}\right\}$$

for which the integrations are to be taken over the effective electrical length $\frac{\pi}{2}\frac{\lambda}{\lambda'}$ corresponding to the actual line length l of the horizontal antenna.

The apparent effective values of formulas (2) confirm, however, only the proper oscillation constant, but generally not the surge impedance, altho a sinusoidal resonance current, flowing in the artificial antenna, would produce the correct effective value as measured at the interference point of the actual antenna, since for such an excitation the resonance current is practically determined by the effective resistance of the combination. The deductions for the apparent effective values are found by means of the derivations for the average values of the antenna potential and antenna current at a certain moment and the relations

$$\left(\begin{matrix}\text{Electric Charge}\\ \text{on Antenna}\end{matrix}\right)=(\text{Average Voltage})\text{ times }(\text{Static Capacity})$$

(Magnetic Flux about Antenna) = (Average Current) times (Static Inductance)

From the figure, it is evident that the correct effective antenna constants vary within the limits

$$\left.\begin{aligned} C_{e_1}{}' &= [81\% \text{ to } 100\%] \text{ static value} \\ L_{e_1}{}' &= [50\% \text{ to } 33\%] \text{ static value} \\ r_e{}' &= [50\% \text{ to } 33\%] \text{ static value} \end{aligned}\right\} \begin{array}{l}\text{from unloaded to} \\ \text{coil-loaded state}\end{array}$$

The limit for the unloaded condition confirms Dr. Miller's results, which are readily found from formulas (1) by equating λ' to λ.

The other limit is based on the fact that for a large loading inductance the effective potential varies but little along the antenna and the current decreases almost linearly towards the open end, for which distributions

$$C_{e_1}{}' = \frac{\left[\int_o^l C\,dx\right]^2}{\int_o^l C\,dx} = l\,C = C_A = \text{static value}$$

$$L_{e_1}{}' = \int_o^l L\left(1-\frac{x}{l}\right)^2 dx = \frac{1}{3}\,l\,L = \frac{1}{3}\,L_A = 0.33 \text{ static value}$$

$$r_e{}' = \int_o^l r\left(1-\frac{x}{l}\right)^2 dx = 0.33 \text{ static value}$$

If now the oscillation constants for the correct and apparent effective values are compared, one has

$$C_{e_1}{}'\,[L_{e_1}{}' + L_o] = C_{e_2}{}'\,[L_{e_2}{}' + L_o]$$

or

$$A_1\,C_A\,[B_1\,L_A + L_o] = A_2\,C_A\,[B_2\,L_A + L_o]$$

where

$$A_1 = \frac{8\,\lambda'}{\pi\,\lambda}\;\frac{sin^2\,90\,\frac{\lambda}{\lambda'}}{\pi\,\frac{\lambda}{\lambda'} + sin\,180\left[1-\frac{\lambda}{\lambda'}\right]}$$

$$B_1 = \frac{\lambda'}{2\,\pi\,\lambda}\;\frac{\pi\,\frac{\lambda}{\lambda'} - sin\,180\left[1-\frac{\lambda}{\lambda'}\right]}{sin^2\,90\,\frac{\lambda}{\lambda'}}$$

$$A_2 = \frac{2\,\lambda'}{\pi\,\lambda}\;sin\,90\,\frac{\lambda}{\lambda'}$$

$$B_2 = \frac{2\,\lambda'}{\pi\,\lambda}\;\frac{\left[1-cos\,90\,\frac{\lambda}{\lambda'}\right]}{sin\,90\,\frac{\lambda}{\lambda'}}$$

The static self-inductance L_A of the antenna can therefore be calculated from the formula

$$L_A{}^H = \frac{A_1 - A_2}{A_2 B_2 - A_1 B_1} \cdot L_o{}^H$$

If for the unloaded antenna $\lambda = 685$ m., and for the same antenna with an inserted loading and inductance, $L_o = 0.000246$ henry, the resonance wave length $\lambda' = 1{,}015$ m., is measured, the correct effective constants become

$$\left.\begin{array}{l} C_{e_1}' = 0.967\, C_A \\ L_{e_1} = 0.393 L_A \end{array}\right\} \text{ hence } \left\{\begin{array}{l} A_1 = 0.967 \\ B_1 = 0.393 \end{array}\right.$$

The apparent effective values lead to

$$\left.\begin{array}{l} C_{e_2}' = 0.826\, C_A \\ L_{e_2}' = 0.56\, L_A \end{array}\right\} \text{ hence } \left\{\begin{array}{l} A_2 = 0.826 \\ B_2 = 0.56 \end{array}\right.$$

The static value of the inductance is then

$$L_A = \frac{0.967 - 0.826}{0.826 \times 0.56 - 0.967 \times 0.393}\, 0.000246 = 0.000413 \text{ henry}$$

The static line capacity can be calculated from the expression

$$C_A{}^F = \frac{\lambda^2}{144 \times 10^{16}\, L_A{}^H}$$

since for the unloaded antenna

$$\lambda = 6\,\pi \times 10^8 \sqrt{C_e{}^F \cdot L_e{}^H} = 6\,\pi \times 10^8 \sqrt{\left[\frac{8}{\pi^2} C_A\right] \frac{L_A}{2}}$$
$$= 12 \times 10^8 \sqrt{C_A\, L_A}$$

where λ is expressed in meters.

Hence

$$C_A = \frac{685^2}{144 \times 10^{16} \times 413 \times 10^{-16}} = 0.00079 \text{ microfarad.}$$

The correct effective antenna constants are therefore

$$C_{e_1}' = 0.967 \times 0.00079 = 0.000764 \text{ microfarad}$$
$$L_{e_1}' = 0.393 \times 0.000413 = 0.0001625 \text{ henry}$$

If these values are inserted in the Thomson formula

$$\lambda' = 6\,\pi \times 10^8 \sqrt{C_{e1}' \left[L_{e1}' + L_o\right]}$$
$$= 6 \times 10^8\, \pi \sqrt{\frac{0.764}{10^9} \left[0.0001625 + 0.000246\right]}$$
$$= 1{,}050 \text{ m. approximately,}$$

a value is obtained which differs about 3.35 per cent from the measured wave length, $\lambda' = 1{,}015$ m. This discrepancy is probably due to the fact that the inserted loading inductance has been calibrated with an audio current.

Moreover, according to the above considerations the effective loading self induction L_o can be calculated from the formula

$$L_o^H = \frac{A_2 B_2 - A_1 B_1}{A_1 - A_2} L_A^H$$

Applying this result, for instance, to one of Professor Morecroft's investigations, where the static inductance of the aerial is $L_A = 0.375$ henry, the fundamental frequency $f = 100.5$ cycles per second for the unloaded antenna, and $f' = 60.2$ cycles per second for the same antenna but for an inserted inductance at the ground side, gives

$$\frac{\lambda}{\lambda'} = \frac{f'}{f} = 0.599 \qquad A_1 = 0.982 \qquad A_2 = 0.860$$
$$B_1 = 0.377 \qquad B_2 = 0.5415$$

hence

$$L_o = \frac{0.860 \times 0.541 - 0.982 \times 0.377}{0.982 - 0.860} 0.375 = 0.295 \text{ henry},$$

which is in very close approximation the measured value $L_o = 0.292$ henry in Professor Morecroft's experiments.

SUMMARY: Formulas for the correct and the effective inductance, capacity, and resistance of an antenna are given. The relation of the values given to the corresponding static values is studied. The results obtained are shown to agree closely with the experimentally determined values for a coil-loaded antenna.

FURTHER DISCUSSION ON

"THE USE OF GROUND WIRES AT REMOTE CONTROL STATIONS"

BY

LIEUTENANT-COMMANDER A. HOYT TAYLOR AND LIEUTENANT A. CROSSLEY

Ellery W. Stone (by letter): The above paper and Mr. Taylor's preceding papers on the use of ground and water wires have been of great interest to me, particularly because of the account of the elimination of interference from local transmitters at control stations. In this connection, a description of the conditions which obtained at the Naval Radio Station at San Diego, Cal., during my tour of duty there as District Communication Superintendent, may prove of interest.

As at all naval district center stations, the San Diego station consisted of a receiving and control station and a transmitting station, separated from each other by some 15 miles (24 km.). The control station, at which were also located the administration offices, was situated on Point Loma, a bluff some 500 feet (150 m.) high, overlooking the ocean. The transmitting station was located at Chollas Heights, some distance inland, at about the same elevation. At the control station, three receiving circuits were in operation which controlled respectively a high power arc (200 kw.), a low power arc (30 kw.), and a spark set (10 kw.). In addition, a fourth receiving circuit was in constant use for the interception of dispatches from German stations on both sides of the Atlantic. Altho considerable ingenuity was employed to eliminate the troublesome interference from our own transmitters as well as that from two other local military stations, Fort Rosecrans and the Naval Air Station, it was never found possible to eradicate it completely so long as elevated receiving antennas were employed.

Due to the success of Mr. Taylor's work at Great Lakes and other stations, experiments were carried on at the Naval Air Station on North Island, which is located in San Diego Bay, to see if the adoption of water wires would not eliminate this difficulty. The results, even at the start, were so gratifying that it was found possible to remove the control station to North Island, about five miles (8 km.) nearer our transmitters at Chollas Heights, and to obtain satisfactory, simultaneous operation of all circuits. It is a pleasure, therefore, to testify as to the value of Mr. Taylor's contribution to the advancement of the art.

In his first and third papers, Mr. Taylor accounts for the operation of the buried antennas in the induction of potentials within the wire by the horizontal component of the tilted electrostatic wave front. This does not explain the operation of the submarine antenna, however, for the wave front is perpendicular to the equatorial plane when passing over sea water *and has a zero horizontal component*. Since the most satisfactory operation of the buried antenna was always obtained when it was surrounded with a medium of good conductivity, such as moist soil or water (fresh or salt), in which case there would be very little tilting of the wave front—if any, it appears desirable to suggest a different explanation for the action of antennas of this type.

According to the usually accepted theory of wave propagation, the electrostatic strains from the transmitting antenna to earth become detached at the quarter wave length point and move over the surface of the earth, the static potentials at the earth's surface causing currents to flow in the ground which are accompanied by a magnetic field at right angles to the vertical static lines of force and to the horizontal earth currents.

It must be understood that in the consideration of the phenomenon of wave propagation it is not correct to differentiate between the three forces involved, that is, static, kinetic (conductive), and magnetic, to the extent of saying that the actual *propagation* depends upon either one or two to the exclusion of the remainder. On the contrary, all three are involved, and in the case of reception by the elevated, grounded antenna, all three are used.

With the buried antenna, however, despite the *presence* of the three forces of propagation, the shielding effect of the conductive layer above the wire eliminates any action of the electrostatic space strains, and as a result we appear to make use only of the earth currents and their accompanying electromagnetic field. We may consider that potentials are induced in the antenna wire by the magnetic field surrounding the earth currents and by electrostatic induction from the earth currents. This static induction takes place by virtue of the insulation between the two circuits, the earth and the wire, and is not to be confused with the static space strains above the earth's surface. The latter are made use of in reception by the loop antenna but not with buried wires. Since the earth currents are greatest in media of high conductivity, there will be a maximum magnetic and static induction of potentials with consequent greatest signal strength.

The explanation advanced above is of course not counter to Mr. Taylor's explanations concerning the shielding effect against strays and so on.

San Francisco, California,
July 12, 1920.

ELECTROSTATICALLY COUPLED CIRCUITS*

By

Louis Cohen, Ph.D.

(Consulting Engineer, Washington, District of Columbia)

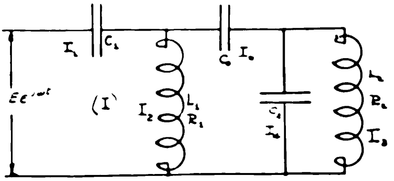
In radio frequency current work, coupled circuits in one form or another are generally employed; coupling between circuits may be either magnetic, direct, electrostatic, or any combination of the above types of coupling.

The magnetic type of coupling has been discussed extensively by a number of authors, and some attention has also been given to the theory of direct coupling, but very little consideration, if any, has been given to the subject of electrostatic coupling.

It is the purpose of this paper to discuss briefly the theory of electrostatic coupling from the point of view of energy transfer and sharpness of tuning, and compare results with these obtained in magnetic coupled circuits.

Electrostatic coupling is commonly understood to mean the electrical linking of two independent circuits by means of a condenser as shown in the diagram, and the term is used here in this sense. In the figures, C_1, L_1, and C_2, L_2 constitute the primary and secondary circuits respectively and C_o the coupling condenser.

To begin with it is desirable to ascertain the oscillation frequency constant of the system, which can be most readily done by determining the frequency for which the reactance of the system vanishes.



Electrostatically Coupled Circuit

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* Received by the Editor, July 15, 1919.

The reactance of C_2 and L_2 in parallel is,

$$\frac{1}{Z_2}=\frac{1}{L_2 j\omega}+C_2 j\omega=\frac{1-L_2C_2\omega^2}{L_2 j\omega},$$

$$Z_2=\frac{L_2 j\omega}{1-L_2C_2\omega^2},$$

Adding the reactance of C_o in series, we have

$$Z'=Z_o+Z_2=\frac{1}{C_o j\omega}+\frac{L_2 j\omega}{1-L_2C_2\omega^2}$$

$$=\frac{1-L_2C_2\omega^2-L_2C_o\omega^2}{C_o j\omega\,(1-L_2C_2\omega^2)}$$

Adding the reactance L_1 in parallel,

$$\frac{1}{Z_1}=\frac{1}{L_1 j\omega}+\frac{1}{Z'}=\frac{1}{L_1 j\omega}+\frac{C_o j\omega\,(1-L_2C_2\omega^2)}{1-L_2C_2\omega^2-L_2C_o\omega^2}$$

$$=\frac{1-L_2C_2\omega^2-L_2C_o\omega^2-L_1C_o\omega^2\,(1-L_2C_2\omega^2)}{L_1 j\omega\,(1-L_2C_2\omega^2-L_2C_o\omega^2)}$$

and

$$Z_1=\frac{L_1 j\omega\,(1-L_2C_2\omega^2-L_2C_o\omega^2)}{1-L_2C_2\omega^2-L_2C_o\omega^2-L_1C_o\omega^2\,(1-L_2C_2\omega^2)}$$

Adding the reactances of C_1 in series, we get the total reactance of the system,

$$Z_t=\frac{1}{C_1 j\omega}+\frac{L_1 j\omega\,(1-L_2C_2\omega^2-L_2C_o\omega^2)}{1-L_2C_2\omega^2-L_2C_o\omega^2-L_1C_o\omega^2+L_1L_2C_oC_2\omega^4}$$

$$Z_t=\frac{1-L_2C_2\omega^2-L_2C_o\omega^2-L_1C_o\omega^2+L_1L_2C_oC_2\omega^4-L_1C_1\omega^2+L_1L_2C_1C_2\omega^4+L_1L_2C_1C_o\omega^4}{C_1 j\omega\,(1-L_2C_2\omega^2-L_2C_o\omega^2-L_1C_o\omega^2+L_1L_2C_oC_2\omega^4)}. \qquad (1)$$

Putting the numerator of equation (1) equal to zero and solving for ω, we get,

$$\omega^2=\frac{(L_2C_2+L_2C_o+L_1C_1+L_1C_o)\pm\sqrt{(L_2C_2+L_2C_o+L_1C_1+L_1C_o)^2-4L_1L_2(C_oC_2+C_1C_2+C_1C_o)}}{2L_1L_2\,(C_oC_2+C_1C_2+C_1C_o)}. \qquad (2)$$

When the circuits are syntonised, $L_1C_1=L_2C_2$, the above equation reduces to,

$$\omega^2=\frac{\left(2+\frac{C_o}{C_1}+\frac{C_o}{C_2}\right)\pm\left(\frac{C_o}{C_1}+\frac{C_o}{C_2}\right)}{2LC\left(1+\frac{C_o}{C_1}+\frac{C_o}{C_2}\right)}. \qquad (3)$$

The expression for ω^2 given by equation (3) gives the value of ω for which the reactance of the system is zero. There are two values of ω for the condition of vanishing reactances, and the system has accordingly two natural frequencies of oscillations which are as follows:

$$
\left.\begin{aligned}
f &= \frac{1}{2\pi\sqrt{LC}},\\
f_1 &= \frac{1}{2\pi\sqrt{LC\left(1+\frac{C_0}{C_1}+\frac{C_0}{C_2}\right)}}.
\end{aligned}\right\} \quad (4)
$$

It will be shown further on that even for tight coupling to get the maximum energy transfer to the secondary circuits, C_0 is very small in comparison with C_1 or C_2. $\frac{C_0}{C}$ is of the order of $\frac{1}{200}$ or less, and the difference between the two frequencies is accordingly very small indeed for all degrees of coupling, so that for practical purposes the system may be considered as a single frequency system.

We shall now consider the current distribution in the circuits so as to determine the current amplitude in the secondary circuit and sharpness of tuning, and compare the results with those of a magnetic coupled system.

Refering to the diagram above, we have the following equations for the current distribution in the circuits:

$$
\left.\begin{aligned}
&\frac{1}{C_1 j\omega} I_1 + (L_1 j\omega + R_1)\, I_2 = E,\\
&\frac{1}{C_1 j\omega} I_1 + \frac{1}{C_0 j\omega} I_0 + (L_2 j\omega + R_2)\, I_3 = E,\\
&\frac{1}{C_1 j\omega} I_1 + \frac{1}{C_0 j\omega} I_0 + \frac{1}{C_2 j\omega} I_4 = E.
\end{aligned}\right\} \quad (5)
$$

Also the supplementary relations,

$$
\left.\begin{aligned}
I_0 &= I_3 + I_4,\\
I_1 &= I_2 + I_0 = I_2 + I_3 + I_4.
\end{aligned}\right\} \quad (6)
$$

Substituting the values of I_0 and I_1 from (6) into (5), we get,

$$
\left.\begin{aligned}
&\left(\frac{1}{C_1 j\omega} + L_1 j\omega + R_1\right) I_2 + \frac{1}{C_1 j\omega} I_3 + \frac{1}{C_1 j\omega} I_4 = E,\\
&\frac{1}{C_1 j\omega} I_2 + \left(\frac{1}{C_1 j\omega} + \frac{1}{C_0 j\omega} + L_2 j\omega + R_2\right) I_3 +\\
&\qquad\qquad \left(\frac{1}{C_1 j\omega} + \frac{1}{C_0 j\omega}\right) I_4 = E,\\
&\frac{1}{C_1 j\omega} I_2 + \left(\frac{1}{C_1 j\omega} + \frac{1}{C_0 j\omega}\right) I_3 + \left(\frac{1}{C_1 j\omega} +\right.\\
&\qquad\qquad \left.\frac{1}{C_0 j\omega} + \frac{1}{C_2 j\omega}\right) I_4 = E.
\end{aligned}\right\} \quad (7)
$$

Eliminating I_2 from equations (7), we get the following:

$$\left\{\frac{1}{C_1^2\,\omega^2}+\left(\frac{1}{C_1 j\omega}+L_1 j\,\omega+R_1\right)\left(\frac{1}{C_1 j\,\omega}+L_2 j\,\omega+R_2+\frac{1}{C_o j\,\omega}\right)\right\} I_3$$
$$+\left\{\frac{1}{C_1^2\,\omega^2}+\left(\frac{1}{C_1 j\,\omega}+\frac{1}{C_o j\,\omega}\right)\left(\frac{1}{C_1 j\,\omega}+L_1 j\,\omega+R_1\right)\right\} I_4$$
$$=E\,(L_1 j\,\omega+R_1)\;(L_2 j\,\omega+R_2)\,I_3-\frac{1}{C_2 j\,\omega}\,I_4=0 \qquad (8)$$

In the first equation of (8) we may, to a high degree of approximation, neglect the terms $\frac{1}{C_1 j\,\omega}+L_2 j\,\omega+R_2$ in comparison with the term $\frac{1}{C_o j\,\omega}$ in the factor of coefficient of I_3. The terms neglected are very small indeed in comparison with $\frac{1}{C_o j\,\omega}$ so no appreciable error is introduced thereby and the subsequent work is very much simplified by it. With the assumption indicated above, and multiplying thruout by $C_1 j\,\omega$ the first equation (8) takes the form,

$$\left\{-\frac{1}{C_1 j\,\omega}+\frac{1-L_1 C_1\omega^2}{C_o j\,\omega}+R_1\frac{C_1}{C_o}\right\} I_3+\left\{-\frac{1}{C_1 j\,\omega}+\frac{1-L_1 C_1\,\omega^2}{C_1 j\,\omega}+\right.$$
$$\left.\frac{1-L_1 C_1\,\omega^2}{C_o j\,\omega}+R_1\frac{C_1}{C_o}\right\} I_4=E\,(-L_1\,C_1\,\omega^2+R_1\,C_1 j\,\omega) \qquad (9)$$

From the second equation of (8), we have

$$I_3=\frac{I_4}{-L_2\,C_2\,\omega^2+R_2\,C_2 j\,\omega}. \qquad (10)$$

Introducing the value of I_3 from (10) into (9) and simplifying, we obtain the following:

$$\left\{\left(-\frac{1}{C_1 j\omega}+\frac{1-L_1C_1\omega^2}{C_o j\omega}+R_1\frac{C_1}{C_o}\right)(1-L_2C_2\omega^2+R_2C_2j\omega)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)\right\}I_4= E(-L_1C_1\omega^2+R_1C_1j\omega)(-L_2C_2j\omega+R_2C_2j\omega), \qquad (11)$$

and

$$I_4=\frac{E\{(-L_1C_1\omega^2+R_1C_1j\omega)(-L_2C_2\omega^2+R_2C_2j\omega)}{\left(-\frac{1}{C_1j\omega}+\frac{1-L_1C_1\omega^2}{C_oj\omega}+R_1\frac{C_1}{C_o}\right)\left(1-L_2C_2\omega^2+R_2C_2j\omega\right)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)} \qquad (12)$$

The current in the inductance coil of the secondary circuit is,

$$I_3=\frac{E(-L_1C_1\omega^2+R_1C_1j\omega)}{\left(-\frac{1}{C_1j\omega}+\frac{1-L_1C_1\omega^2}{C_oj\omega}+R_1\frac{C_1}{C_o}\right)(1-L_2C_2\omega^2+R_2C_2j\omega)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)}. \qquad (13)$$

For resonance condition, $L_1C_1\omega^2=L_2C_2\omega^2=1$, equations (12) and (13) reduce to,

$$\left.\begin{aligned} I_4&=\frac{E(-1+R_1C_1j\omega)(-1+R_2C_2j\omega)}{\left(-\frac{1}{C_1j\omega}+R_1\frac{C_1}{C_o}\right)R_2C_2j\omega}, \\ I_3&=\frac{E(-1+R_1C_1j\omega)}{\left(-\frac{1}{C_1j\omega}+R_1\frac{C_1}{C_o}\right)R_2C_2j\omega}. \end{aligned}\right\} \qquad (14)$$

In practical radio circuits, $R_1C_1j\omega$ and $R_2C_2j\phi\omega$ are of the order of magnitude of 0.01 or 0.02, and hence may be neglected in comparison with unity, so we finally have for the currents in the secondary circuit at resonance condition,

$$\left.\begin{aligned} I_4 &= \frac{E}{R_1 R_2 \dfrac{C_1 C^2}{C_o} j\omega - R_2 \dfrac{C_2}{C_1}}, \\ I_3 &= \frac{-E}{R_1 R_2 \dfrac{C_1 C_2}{C_o} j\omega - R_2 \dfrac{C_2}{C_1}}. \end{aligned}\right\} \qquad (15)$$

The absolute values of the currents are,

$$\left.\begin{aligned} \bar{I}_4 &= \frac{E}{\sqrt{\left(R_1 R_2 \dfrac{C_1 C_2 \omega}{C_o}\right)^2 + \left(R_2 \dfrac{C_2}{C_1}\right)^2}}, \\ \bar{I}_3 &= \frac{-E}{\sqrt{\left(R_1 R_2 \dfrac{C_1 C_2 \omega}{C_o}\right)^2 + \left(R_2 \dfrac{C_2}{C_1}\right)^2}}. \end{aligned}\right\} \qquad (16)$$

MAGNETICALLY COUPLED CIRCUITS. For magnetic coupling we have the following well-known expression for the current in the secondary circuit:

$$I_2 = \frac{E M j \omega}{\left(L_1 j\omega + \dfrac{1}{C_1 j\omega} + R_1\right)\left(L_2 j\omega + \dfrac{1}{C_2 j\omega} + R_2\right) + M^2 \omega^2}. \qquad (17)$$

where M is the mutual inductance between the circuits and E is the amplitude of impressed emf. on primary circuit. Equation (17) may be put in the following form:

$$I_2 = \frac{E M j \omega}{\left(\dfrac{1 - L_1 C_1 \omega^2}{C_1 j\omega} + R_1\right)\left(\dfrac{1 - L_2 C_2 \omega^2}{C_2 j\omega} + R_2\right) + M^2 \omega^2}. \qquad (18)$$

For resonance conditions, $L_1 C_1 \omega^2 = L_2 C_2 \omega^2 = 1$, equation (18) reduces to,

$$I_2 = \frac{E M j \omega}{R_1 R_2 + M^2 \omega^2}. \qquad (19)$$

The current in the secondary circuit is a maximum when $M^2 \omega^2 = R_1 R_2$, and in that case

$$I_2 = \frac{E}{2\sqrt{R_1 R_2}}. \qquad (20)$$

The relative merits of the two types of coupling, from the standpoint of secondary curient and sharpness of tuning, can be best shown by a numerical example using the data of practical radio circuits and comparing results obtained by substituting in equations (16) and (20); and equations (13) and (18).

Example:

$R_1 = R_2 = 25$ ohms,
$C_1 = C_2 = 0.001$ microfarad,
$C_n = 0.000005$ microfarad,
$\omega = 3 \times 10^5$.

By (16), resonance current, electrostatic coupling,

$$I = \frac{E}{\sqrt{(37.5)^2 + (25)^2}} = \frac{E}{45} \text{ approximately.}$$

By (20), resonance current, magnetic coupling,

$$I = \frac{E}{50}.$$

The current amplitudes are of the same order of magnitude in both cases.

SHARPNESS OF TUNING. The following table, giving the calculated values of the current by equations (13) and (18) for several frequencies on either side of the resonance frequency, shows that the static coupling gives a sharper resonance effect.

ω	I_3 By Equation (13)	I_2 By Equation (18)
$0.90 \times 3 \times 10^5$	$0.31 \times 10^{-4} E$	$0.45 \times 10^{-4} E$
$0.94 \times 3 \times 10^5$	$0.96 \times 10^{-4} E$	$1.40 \times 10^{-4} E$
$0.98 \times 3 \times 10^5$	$10.00 \times 10^{-4} E$	$12.00 \times 10^{-4} E$
$1.00 \times 3 \times 10^5$	$22.20 \times 10^{-4} E$	$20.00 \times 10^{-4} E$
$1.02 \times 3 \times 10^5$	$8.00 \times 10^{-4} E$	$14.60 \times 10^{-4} E$
$1.06 \times 3 \times 10^5$	$1.10 \times 10^{-4} E$	$1.86 \times 10^{-4} E$
$1.10 \times 3 \times 10^5$	$0.44 \times 10^{-4} E$	$0.68 \times 10^{-4} E$

SUMMARY: The expressions for the secondary frequencies secondary current, and secondary resonance current of electrostatically coupled circuits are derived theoretically. The system is shown to be practically monofrequent even for fairly close coupling between primary and secondary circuits.

The results obtained are compared with the corresponding expressions for inductive coupling. In a practical numerical example, it is found that the secondary resonance currents in the two cases are of the same order of magnitude but that the electrostatic coupling yields superior sharpness of resonance.

THE WAVE LENGTH RELATION FOR A GENERALIZED BESSEL'S ANTENNA*

By

A. Press

(Assistant Professor of Electrical Engineering, University of California, Berkeley, California)

In the December issue for 1918 of the Proceedings of The Institute of Radio Engineers the writer indicated how the problem of a vertical antenna should be treated, taking into account the variable distribution of both inductance and capitance per unit of length of the aerial. The subjoined paper goes into the matter more fully and a plot is given for undamped waves both of the voltage and current distributions for such an antenna. An antenna length (height) l is assumed, together with an impressed frequency f, such that for a wave length λ the following relation is assumed to hold:

$$\frac{\pi l}{\lambda} = 1.$$

It is then found that a voltage nodal point exists near the bottom of the antenna which corresponds to a current antinode. Thus, when adjustments are made along the coupling coil to obtain maximum current in the antenna wire, this corresponds to a maximum potential with respect to neutral of the free end of the aerial In the analysis below no account is taken of resistance. Where radiation takes place, the voltage and current distributions will occur as before, but the voltage will be practically in time phase with the current. A true fixed antinode is, of course, a physical impossibility just as much so as in the vibration of a violin string.

Assuming that the antenna is $l=100$ feet (30.5 m.) in height, then the entire horizontal scale $\left(\frac{x}{l}\right)$, Figure 2 corresponds to unity, for this height. If, then, at the top of the antenna a maximum potential of 100,000 volts with respect to neutral is assumed, this would mean from the curve that at 27.5 feet (8.4 m.) the voltage would be one-tenth of the maximum voltage. The nodal point, however, occurs at 20 feet (6 m.), so that for this adjustment maximum potential magnification will occur at the free end of the aerial. Correspondingly for this point the aerial current will be a maximum.

---

* Received by the Editor, June 4, 1919.

A vertical antenna is assumed with negligible resistivity and leakage conductance, and then the differential equations to be solved are

$$L\frac{di}{dt} = -\frac{dv}{dx}$$

$$C\frac{dv}{dt} = -\frac{di}{dx}$$

where v and i are the voltage and current functions with respect to x and t. It has been shown by the writer (previous citation) that as a function of x up to 200 feet (61 m.) it is possible to write

$$L = a\,x^{n}$$

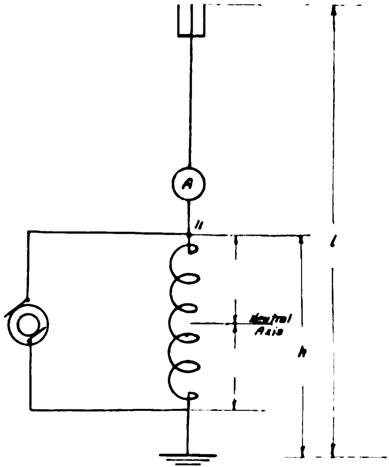

FIGURE 1

where $$a = 7 \times 10^{-9};\quad n = 0.13$$

provided L is measured in henrys per centimeter. Moreover, it has also been shown that with the same degree of approximation

$$C = \beta\,x^{-n}$$

where if C is measured in farads per centimeter

$$\beta = 1.58 \times 10^{-12}$$

and n is as before.

Separating out the variables the following equations were deduced (previous citation):

$$\frac{d^2 i}{d x^2} + \frac{n}{x}\frac{d i}{d x} = q^2 i$$

$$\frac{d^2 v}{d x^2} - \frac{n}{x}\frac{d v}{d x} = q^2 v$$

with

$$q^2 = -\frac{1}{V^2}\frac{d^2}{d t^2}$$

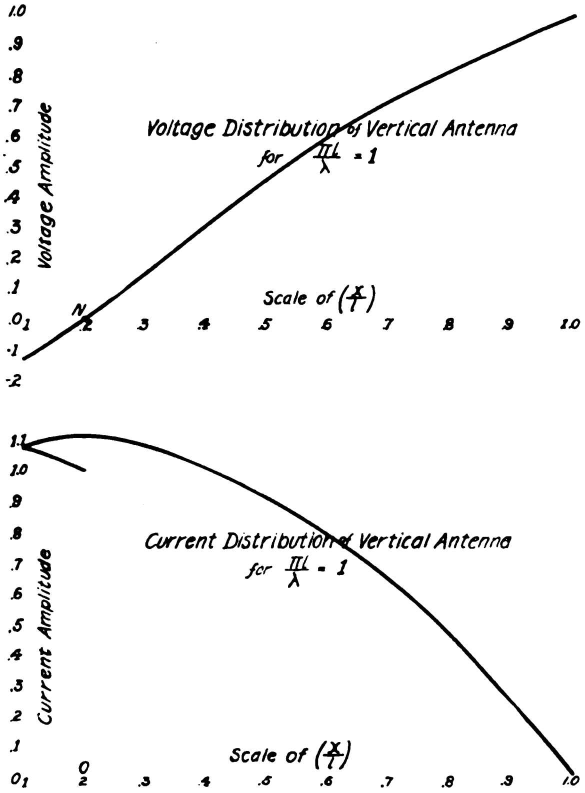

FIGURE 2.

and V is the velocity of light in centimeters. To meet the conditions of the problem the complete solution

$$i = A \cdot \frac{I_m(q x)}{x^m} + B \cdot \frac{I_{-m}(q x)}{x^m}$$

is to be investigated subject to the condition that

$$m = \tfrac{1}{2}(n-1) = \tfrac{1}{2}(0.13-1) = -0.435.$$

This will mean that all the x's of the term I_{-m} will have plus indices.

To satisfy the boundary conditions we note that at

$$x = l = \text{antenna height},$$

the current must be zero. Again, at $x = h$, where the exciting circuit is coupled to the antenna, the voltage to neutral is assumed to be

$$v_h = V/_{x=h} = E_h \sin pt.$$

Introducing the first condition we have

$$0 = A \cdot I_m(ql) + B \cdot I_{-m}(ql)$$

$$B = -A \cdot \frac{I_m(ql)}{I_{-m}(ql)}.$$

For the voltage relation again, we must have

$$C\frac{dv}{dt} = -\frac{di}{dx}$$

However, for Bessel's Functions, we also have

$$\frac{d}{dx} \cdot \frac{I_m}{x^m} = q \cdot \frac{I_{m+1}}{x^m}$$

$$\frac{d}{dx} \cdot \frac{I_{-m}}{x^m} = q \cdot \frac{I_{-m-1}}{x^m}$$

(see Heaviside "Electromagnetic Theory," volume 2, page 245) and thus

$$C\frac{dv}{dt} = -Aq \cdot \frac{I_{m+1}}{x^m} - Bq \cdot \frac{I_{-m-1}}{x^m}$$

or again

$$v = \left\{ I_m(ql) \cdot I_{-m-1}(qx) - I_{-m}(ql) \cdot I_{m+1}(qx) \right\} \cdot x^{m+1} \cdot \frac{Aq}{t_1 \beta I_{-m}(ql)}.$$

This voltage equation must conform to the condition for $x = h$. However, it will be best to consider the maximum potential at $x = l$, and then

$$v_l = \left\{ I_m \cdot I_{-m-1} - I_{-m} \cdot I_{m+1} \right\}_{ql} \cdot l^{m+1} \cdot \frac{Aq}{t_1 \beta I_{-m}(ql)} = E \sin pt$$

from which A/t_1 can be evaluated. Thus we have on substitution

$$v = \left\{ \frac{I_m(ql) \cdot I_{-m-1}(qx) - I_{-m}(ql) \cdot I_{m+1}(qx)}{I_m(ql) \cdot I_{-m-1}(ql) - I_m(ql) \cdot I_{m+1}(ql)} \right\} \cdot \left(\frac{x}{l}\right)^{m+1} E \sin pt$$

Similarly, for the current relation we also have

$$i = \left\{ \frac{I_{-m}(ql) \cdot I_m(qx) - I_m(ql) \cdot I_{-m}(qx)}{\text{same denominator}} \right\} \frac{\beta}{(xl)^m} \cdot \frac{p}{ql} E \cos pt$$

The denominator occurring in both the v and i functions can be very much simplified by means of the relation respecting conjugate functions given in Heaviside, "Electromagnetic Theory," volume 2, page 245. A text dealing with the "Symbolic Methods of Oliver Heaviside" is in course of preparation. Moreover, by introducing the P_{-m} functions deduced from the I_m functions by dividing out with the first term, the following simplified formulas result:

$$v = \left\{ P_m(ql) \cdot P_{-m}(qx) - \left(\frac{\pi x}{\lambda}\right)^2 \left(\frac{x}{l}\right)^m \cdot \frac{P_{-m}(ql) \cdot P_{m+1}(qx)}{m(m+1)} \right\} \cdot E \sin pt$$

$$i = \left\{ P_m(ql) \cdot P_{-m}(qx) \left(\frac{l}{x}\right)^{2m} - P_{-m}(ql) \cdot P_m(qx) \right\} \frac{\beta p}{2 m l^{2m}} \cdot E \cos pt$$

The last two formulas are especially calculable when the following general formula for P_m is used:

$$P_m(qx) = 1 + \frac{X}{1(m+1)}\left(1 + \frac{X}{2(m+2)}\left(1 + \frac{X}{3(m+3)}\left(1 + \cdots\right.\right.\right.$$

where in our case

$$X^{\frac{1}{2}} = \frac{\pi x}{\lambda} j$$

The P_m functions, Figure 3, for $m = -0.435$ have been plotted. It should be noted that if for simplicity we take

$$\frac{\pi l}{\lambda} = 1 \quad \text{then} \quad \frac{x}{l} = \frac{\pi x}{\lambda}$$

and therefore only a single argument need be used for investigational purposes. The appended curves for v and i as functions of $\frac{x}{l}$ have been plotted by means of the above formulas.

It will be noticed from the graph that if we consider 80 per cent of the length of the antenna height as measuring ¼ of the wave length, then, since $\pi l = \lambda$,

$$4 \times 0.8 l = 3.2 l \quad \lambda.$$

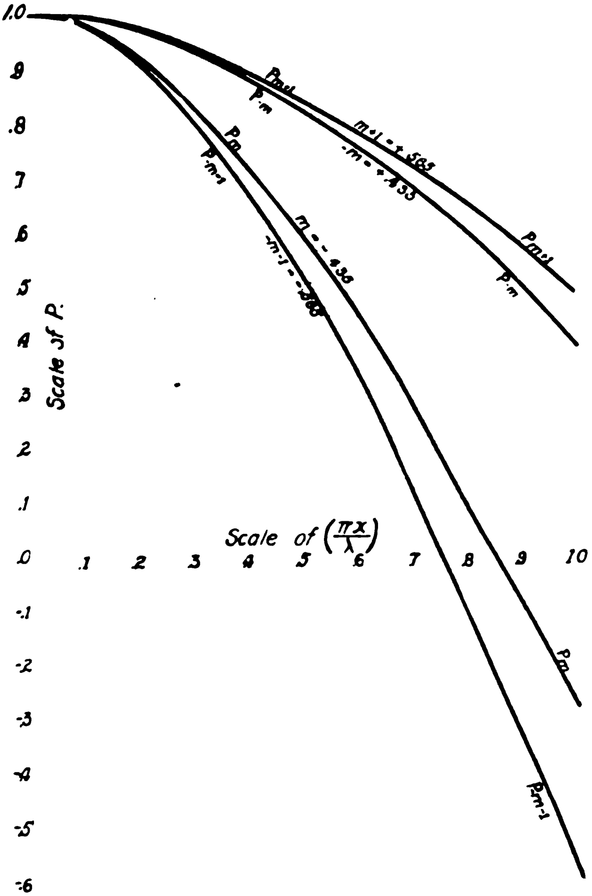

FIGURE 3

Thus, on this basis, the ordinary four times rule substantially holds, but reference should be had to measurements from the true nodal or resonant point along the antenna.

Berkeley, California,
May 29, 1919.

SUMMARY: The differential equations for an antenna of negligible resistance and leakage conductance are set up. Proceeding from previously deduced values of the inductance and capacity per centimeter along the antenna, the solution is obtained in terms of Bessel's Functions for the voltage and current distribution along the antenna. After simplification, the solution is interpreted, and curves enabling its ready employment are given.

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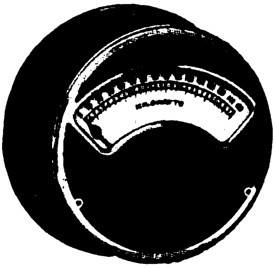

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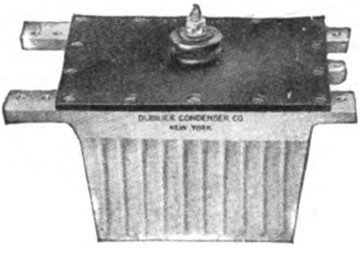

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Volume 8 DECEMBER, 1920 Number 6

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PUBLISHED BY
THE INSTITUTE OF RADIO ENGINEERS, INC.
THE COLLEGE OF THE CITY OF NEW YORK

EDITED BY
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MULTIPLEX RADIO TELEGRAPHY AND TELEPHONY*

By

Francis M. Ryan, J. R. Tolmie, and Roy O. Bach

(Formerly at the University of Washington, Seattle, Washington)

Various methods have been described for the simultaneous reception with a single antenna of radio signals from more than one transmitter. Similar methods have been recorded for multiplex transmission.† These methods of multiplexing may be classified as follows:

(1) Those employing several radio frequencies and obtaining selectivity by means of radio frequency tuning.

(2) Methods using a single radio frequency and several tone frequencies, selectivity being obtained by audio frequency tuning.

(3) The use of a single radio frequency modulated by several intermediate frequencies each of which has been modulated by an audio frequency. Selectivity is obtained in this case by tuning at the intermediate frequencies.

The principal example of (1) is the divided antenna method employed by Marconi for both transmission and reception and which may be used for either telegraphy or telephony. Method (2) has been used by the Telefunken Company,[1] but has never proven satisfactory owing to difficulties arising from strays. Obviously this method is not applicable to telephony. The third method was originated by Mr. R. A. Heising[2] and is applicable to both telegraphy and telephony. Altho somewhat more complicated than method (1), it is superior in so far as secrecy of transmission is concerned. A consideration of the effect of modulation shows that this method utilizes a considerable range of radio frequencies.

---

* Received by the Editor, January 9, 1920. Presented before the Seattle Section of the Institute, October 15, 1919; presented before the Institute at New York, February 4, 1920.

† Since the preparation of this paper, it has come to the writer's attention that John Stone Stone patented in 1905 circuits similar to those described for multiplex transmission and reception.

[1] Zenneck, "Wireless Telegraphy," page 325.

[2] Craft and Colpitts, "Radio Telephony," "Proceedings of the American Institute of Electrical Engineers," March, 1919.

system has accordingly two natural frequencies of oscillations which are as follows:

$$\left.\begin{aligned} f_1 &= \frac{1}{2\pi\sqrt{L\,C}}, \\ f_2 &= \frac{1}{2\pi\sqrt{L\,C\left(1+\frac{C_o}{C_1}+\frac{C_o}{C_2}\right)}}. \end{aligned}\right\} \quad (4)$$

It will be shown further on that even for tight coupling to get the maximum energy transfer to the secondary circuits, C_o is very small in comparison with C_1 or C_2, $\frac{C_o}{C}$ is of the order of $\frac{1}{200}$ or less, and the difference between the two frequencies is accordingly very small indeed for all degrees of coupling, so that for practical purposes the system may be considered as a single frequency system.

We shall now consider the current distribution in the circuits so as to determine the current amplitude in the secondary circuit and sharpness of tuning, and compare the results with those of a magnetic coupled system.

Refering to the diagram above, we have the following equations for the current distribution in the circuits:

$$\left.\begin{aligned} &\frac{1}{C_1 j\,\omega} I_1 + (L_1 j\,\omega + R_1)\, I_2 = E. \\ &\frac{1}{C_1 j\,\omega} I_1 + \frac{1}{C_o j\,\omega} I_o + (L_2 j\,\omega + R_2)\, I_3 = E. \\ &\frac{1}{C_1 j\,\omega} I_1 + \frac{1}{C_o j\,\omega} I_o + \frac{1}{C_2 j\,\omega} I_4 = E. \end{aligned}\right\} \quad (5)$$

Also the supplementary relations,

$$\left.\begin{aligned} I_o &= I_3 + I_4. \\ I_1 &= I_2 + I_o = I_2 + I_3 + I_4. \end{aligned}\right\} \quad (6)$$

Substituting the values of I_o and I_1 from (6) into (5), we get,

$$\left.\begin{aligned} &\left(\frac{1}{C_1 j\,\omega} + L_1 j\,\omega + R_1\right) I_2 + \frac{1}{C_1 j\,\omega} I_3 + \frac{1}{C_1 j\,\omega} I_4 = E. \\ &\frac{1}{C_1 j\,\omega} I_2 + \left(\frac{1}{C_1 j\,\omega} + \frac{1}{C_o j\,\omega} + L_2 j\,\omega + R_2\right) I_3 + \\ &\qquad\qquad \left(\frac{1}{C_1 j\,\omega} + \frac{1}{C_o j\,\omega}\right) I_4 = E. \\ &\frac{1}{C_1 j\,\omega} I_2 + \left(\frac{1}{C_1 j\,\omega} + \frac{1}{C_o j\,\omega}\right) I_3 + \left(\frac{1}{C_1 j\,\omega} + \right. \\ &\qquad\qquad \left. \frac{1}{C_o j\,\omega} + \frac{1}{C_2 j\,\omega}\right) I_4 = E. \end{aligned}\right\} \quad (7)$$

Eliminating I_2 from equations (7), we get the following:

$$\left\{\frac{1}{C_1^2\,\omega^2}+\left(\frac{1}{C_1\,j\omega}+L_1\,j\,\omega+R_1\right)\left(\frac{1}{C_1\,j\,\omega}+L_2\,j\,\omega+R_2+\frac{1}{C_o\,j\,\omega}\right)\right\}I_3$$
$$+\left\{\frac{1}{C_1^2\,\omega^2}+\left(\frac{1}{C_1\,j\,\omega}+\frac{1}{C_o\,j\,\omega}\right)\left(\frac{1}{C_1\,j\,\omega}+L_1\,j\,\omega+R_1\right)\right\}I_4$$
$$=E\,(L_1\,j\,\omega+R_1)\,(L_2\,j\,\omega+R_2)\,I_3-\frac{1}{C_2\,j\,\omega}\,I_4=0 \qquad (8)$$

In the first equation of (8) we may, to a high degree of approximation, neglect the terms $\frac{1}{C_1\,j\,\omega}+L_2\,j\,\omega+R_2$ in comparison with the term $\frac{1}{C_o\,j\,\omega}$ in the factor of coefficient of I_3. The terms neglected are very small indeed in comparison with $\frac{1}{C_o\,j\,\omega}$ so no appreciable error is introduced thereby and the subsequent work is very much simplified by it. With the assumption indicated above, and multiplying thruout by $C_1\,j\,\omega$ the first equation (8) takes the form,

$$\left\{-\frac{1}{C_1\,j\,\omega}+\frac{1-L_1C_1\omega^2}{C_o\,j\,\omega}+R_1\frac{C_1}{C_o}\right\}I_3+\left\{-\frac{1}{C_1\,j\,\omega}+\frac{1-L_1C_1\,\omega^2}{C_1\,j\,\omega}+\right.$$
$$\left.\frac{1-L_1\,C_1\,\omega^2}{C_o\,j\,\omega}+R_1\frac{C_1}{C_o}\right\}I_4=E\,(-L_1\,C_1\,\omega^2+R_1\,C_1\,j\,\omega) \qquad (9)$$

From the second equation of (8), we have

$$I_3=\frac{I_4}{-L_2\,C_2\,\omega^2+R_2\,C_2\,j\,\omega}. \qquad (10)$$

Introducing the value of I_3 from (10) into (9) and simplifying, we obtain the following:

$$\left\{\left(-\frac{1}{C_1 j\omega}+\frac{1-L_1C_1\omega^2}{C_o j\omega}+R_1\frac{C_1}{C_o}\right)(1-L_2C_2\omega^2+R_2C_2j\omega)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)\right\}I_4=$$
$$E(-L_1C_1\omega^2+R_1C_1j\omega)(-L_2C_2j\omega+R_2C_2j\omega), \qquad (11)$$

and

$$I_4=\frac{E\{(-L_1C_1\omega^2+R_1C_1j\omega)(-L_2C_2\omega^2+R_2C_2j\omega)}{\left(-\frac{1}{C_1j\omega}+\frac{1-L_1C_1\omega^2}{C_oj\omega}+R_1\frac{C_1}{C_o}\right)\left(1-L_2C_2\omega^2+R_2C_2j\omega\right)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)} \qquad (12)$$

The current in the inductance coil of the secondary circuit is,

$$I_3=\frac{E(-L_1C_1\omega^2+R_1C_1j\omega)}{\left(-\frac{1}{C_1j\omega}+\frac{1-L_1C_1\omega^2}{C_oj\omega}+R_1\frac{C_1}{C_o}\right)(1-L_2C_2\omega^2+R_2C_2j\omega)+\frac{1-L_1C_1\omega^2}{C_1j\omega}(-L_2C_2\omega^2+R_2C_2j\omega)}. \qquad (13)$$

For resonance condition, $L_1C_1\omega^2=L_2C_2\omega^2=1$, equations (12) and (13) reduce to,

$$\left.\begin{aligned} I_4&=\frac{E(-1+R_1C_1j\omega)(-1+R_2C_2j\omega)}{\left(-\frac{1}{C_1j\omega}+R_1\frac{C_1}{C_o}\right)R_2C_2j\omega},\\ I_3&=\frac{E(-1+R_1C_1j\omega)}{\left(-\frac{1}{C_1j\omega}+R_1\frac{C_1}{C_o}\right)R_2C_2j\omega}.\end{aligned}\right\} \qquad (14)$$

In practical radio circuits, $R_1C_1j\omega$ and $R_2C_2j\phi\omega$ are of the order of magnitude of 0.01 or 0.02, and hence may be neglected in comparison with unity, so we finally have for the currents in the secondary circuit at resonance condition,

$$\left.\begin{aligned} I_4 &= \frac{E}{R_1 R_2 \frac{C_1 C^2}{C_o} j\omega - R_2 \frac{C_2}{C_1}}, \\ I_3 &= \frac{-E}{R_1 R_2 \frac{C_1 C_2}{C_o} j\omega - R_2 \frac{C_2}{C_1}}. \end{aligned}\right\} \qquad (15)$$

The absolute values of the currents are,

$$\left.\begin{aligned} \bar{I}_4 &= \frac{E}{\sqrt{\left(R_1 R_2 \frac{C_1 C_2 \omega}{C_o}\right)^2 + \left(R_2 \frac{C_2}{C_1}\right)^2}}, \\ \bar{I}_3 &= \frac{-E}{\sqrt{\left(R_1 R_2 \frac{C_1 C_2 \omega}{C_o}\right)^2 + \left(R_2 \frac{C_2}{C_1}\right)^2}}. \end{aligned}\right\} \qquad (16)$$

MAGNETICALLY COUPLED CIRCUITS. For magnetic coupling we have the following well-known expression for the current in the secondary circuit:

$$I_2 = \frac{E M j \omega}{\left(L_1 j \omega + \frac{1}{C_1 j \omega} + R_1\right)\left(L_2 j \omega + \frac{1}{C_2 j \omega} + R_2\right) + M^2 \omega^2}. \qquad (17)$$

where M is the mutual inductance between the circuits and E is the amplitude of impressed emf. on primary circuit. Equation (17) may be put in the following form:

$$I_2 = \frac{E M j \omega}{\left(\frac{1 - L_1 C_1 \omega^2}{C_1 j \omega} + R_1\right)\left(\frac{1 - L_2 C_2 \omega^2}{C_2 j \omega} + R_2\right) + M^2 \omega^2}. \qquad (18)$$

For resonance conditions, $L_1 C_1 \omega^2 = L_2 C_2 \omega^2 = 1$, equation (18) reduces to,

$$I_2 = \frac{E M j \omega}{R_1 R_2 + M^2 \omega^2}. \qquad (19)$$

The current in the secondary circuit is a maximum when $M^2 \omega^2 = R_1 R_2$, and in that case

$$I_2 = \frac{E}{2 \sqrt{R_1 R_2}}. \qquad (20)$$

The relative merits of the two types of coupling, from the standpoint of secondary current and sharpness of tuning, can be best shown by a numerical example using the data of practical radio circuits and comparing results obtained by substituting in equations (16) and (20); and equations (13) and (18).

Example:

$R_1 = R_2 = 25$ ohms,
$C_1 = C_2 = 0.001$ microfarad,
$C_0 = 0.000005$ microfarad,
$\omega = 3 \times 10^5$.

By (16), resonance current, electrostatic coupling,

$$I = \frac{E}{\sqrt{(37.5)^2 + (25)^2}} = \frac{E}{45} \text{ approximately.}$$

By (20), resonance current, magnetic coupling,

$$I = \frac{E}{50}.$$

The current amplitudes are of the same order of magnitude in both cases.

SHARPNESS OF TUNING. The following table, giving the calculated values of the current by equations (13) and (18) for several frequencies on either side of the resonance frequency, shows that the static coupling gives a sharper resonance effect.

ω	I_3 By Equation (13)	I_2 By Equation (18)
$0.90 \times 3 \times 10^5$	$0.31 \times 10^{-4} E$	$0.45 \times 10^{-4} E$
$0.94 \times 3 \times 10^5$	$0.96 \times 10^{-4} E$	$1.40 \times 10^{-4} E$
$0.98 \times 3 \times 10^5$	$10.00 \times 10^{-4} E$	$12.00 \times 10^{-4} E$
$1.00 \times 3 \times 10^5$	$22.20 \times 10^{-4} E$	$20.00 \times 10^{-4} E$
$1.02 \times 3 \times 10^5$	$8.00 \times 10^{-4} E$	$14.60 \times 10^{-4} E$
$1.06 \times 3 \times 10^5$	$1.10 \times 10^{-4} E$	$1.86 \times 10^{-4} E$
$1.10 \times 3 \times 10^5$	$0.44 \times 10^{-4} E$	$0.68 \times 10^{-4} E$

SUMMARY: The expressions for the secondary frequencies secondary current, and secondary resonance current of electrostatically coupled circuits are derived theoretically. The system is shown to be practically monofrequent even for fairly close coupling between primary and secondary circuits.

The results obtained are compared with the corresponding expressions for inductive coupling. In a practical numerical example, it is found that the secondary resonance currents in the two cases are of the same order of magnitude but that the electrostatic coupling yields superior sharpness of resonance.

THE WAVE LENGTH RELATION FOR A GENERALIZED BESSEL'S ANTENNA*

By

A. PRESS

(ASSISTANT PROFESSOR OF ELECTRICAL ENGINEERING, UNIVERSITY OF CALIFORNIA, BERKELEY, CALIFORNIA)

In the December issue for 1918 of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS the writer indicated how the problem of a vertical antenna should be treated, taking into account the variable distribution of both inductance and capitance per unit of length of the aerial. The subjoined paper goes into the matter more fully and a plot is given for undamped waves both of the voltage and current distributions for such an antenna. An antenna length (height) l is assumed, together with an impressed frequency f, such that for a wave length λ the following relation is assumed to hold:

$$\frac{\pi l}{\lambda} = 1.$$

It is then found that a voltage nodal point exists near the bottom of the antenna which corresponds to a current antinode. Thus, when adjustments are made along the coupling coil to obtain maximum current in the antenna wire, this corresponds to a maximum potential with respect to neutral of the free end of the aerial. In the analysis below no account is taken of resistance. Where radiation takes place, the voltage and current distributions will occur as before, but the voltage will be practically in time phase with the current. A true fixed antinode is, of course, a physical impossibility just as much so as in the vibration of a violin string.

Assuming that the antenna is $l = 100$ feet (30.5 m.) in height, then the entire horizontal scale $\left(\frac{x}{l}\right)$, Figure 2 corresponds to unity, for this height. If, then, at the top of the antenna a maximum potential of 100,000 volts with respect to neutral is assumed, this would mean from the curve that at 27.5 feet (8.4 m.) the voltage would be one-tenth of the maximum voltage. The nodal point, however, occurs at 20 feet (6 m.), so that for this adjustment maximum potential magnification will occur at the free end of the aerial. Correspondingly for this point the aerial current will be a maximum.

---

* Received by the Editor, June 4, 1919.

A vertical antenna is assumed with negligible resistivity and leakage conductance, and then the differential equations to be solved are

$$L\frac{di}{dt} = -\frac{dv}{dx}$$

$$C\frac{dv}{dt} = -\frac{di}{dx}$$

where v and i are the voltage and current functions with respect to x and t. It has been shown by the writer (previous citation) that as a function of x up to 200 feet (61 m.) it is possible to write

$$L = a x^n$$

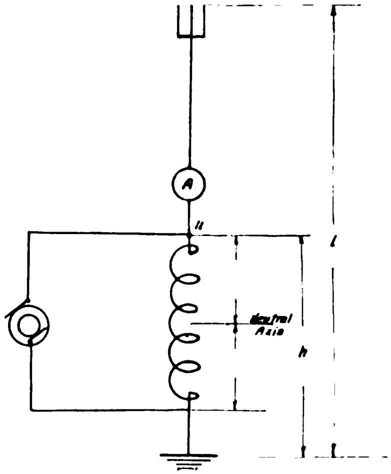

FIGURE 1

where $a = 7 \times 10^{-9}$; $n = 0.13$

provided L is measured in henrys per centimeter. Moreover, it has also been shown that with the same degree of approximation

$$C = \beta x^{-n}$$

where if C is measured in farads per centimeter

$$\beta = 1.58 \times 10^{-12}$$

and n is as before.

Separating out the variables the following equations were deduced (previous citation):

$$\frac{d^2 i}{d x^2} + \frac{n}{x}\frac{d i}{d x} = q^2 i$$

$$\frac{d^2 v}{d x^2} - \frac{n}{x}\frac{d v}{d x} = q^2 v$$

with

$$q^2 = -\frac{1}{V^2}\frac{d^2}{d t^2}$$

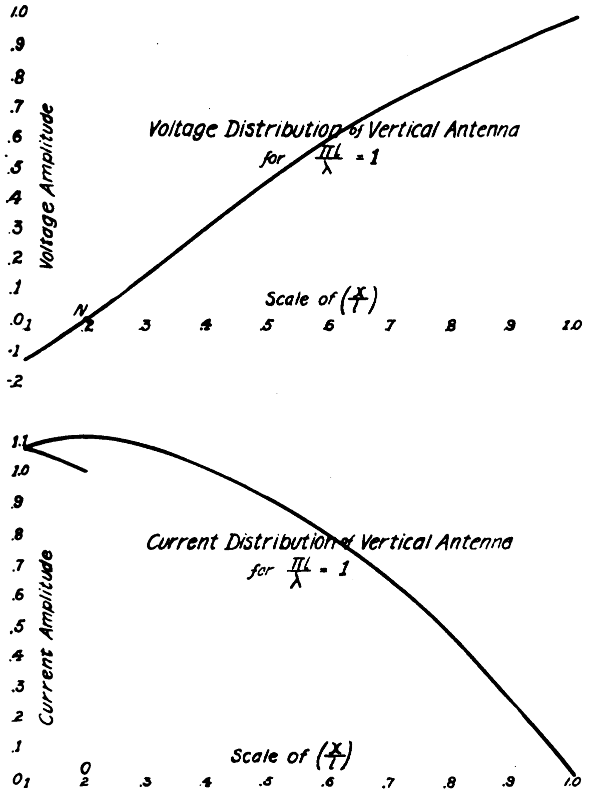

FIGURE 2.

and V is the velocity of light in centimeters. To meet the conditions of the problem the complete solution

$$i = A \cdot \frac{I_m(q x)}{x^m} + B \cdot \frac{I_{-m}(q x)}{x^m}$$

is to be investigated subject to the condition that

$$m = \tfrac{1}{2}(n-1) = \tfrac{1}{2}(0.13-1) = -0.435.$$

This will mean that all the x's of the term I_{-m} will have plus indices.

To satisfy the boundary conditions we note that at

$$x = l = \text{antenna height},$$

the current must be zero. Again, at $x = h$, where the exciting circuit is coupled to the antenna, the voltage to neutral is assumed to be

$$v_h = V/_{x=h} = E_h \sin pt.$$

Introducing the first condition we have

$$0 = A \cdot I_m(ql) + B \cdot I_{-m}(ql)$$

$$B = -A \cdot \frac{I_m(ql)}{I_{-m}(ql)}.$$

For the voltage relation again, we must have

$$C\frac{dv}{dt} = -\frac{di}{dx}$$

However, for Bessel's Functions, we also have

$$\frac{d}{dx} \cdot \frac{I_m}{x^m} = q \cdot \frac{I_{m+1}}{x^m}$$

$$\frac{d}{dx} \cdot \frac{I_{-m}}{x^m} = q \cdot \frac{I_{-m-1}}{x^m}$$

(see Heaviside "Electromagnetic Theory," volume 2, page 245) and thus

$$C\frac{dv}{dt} = -Aq \cdot \frac{I_{m+1}}{x^m} - Bq \cdot \frac{I_{-m-1}}{x^m}$$

or again

$$v = \{I_m(ql) \cdot I_{-m-1}(qx) - I_{-m}(ql) \cdot I_{m+1}(qx)\} \cdot x^{m+1} \cdot \frac{Aq}{t_1 \beta I_{-m}(ql)}.$$

This voltage equation must conform to the condition for $x = h$. However, it will be best to consider the maximum potential at $x = l$, and then

$$v_l = \{I_m \cdot I_{-m-1} - I_{-m} \cdot I_{m+1}\}_{ql} \cdot l^{m+1} \cdot \frac{Aq}{t_1 \beta I_{-m}(ql)} = E \sin pt$$

from which A/t_1 can be evaluated. Thus we have on substitution

$$v = \left\{ \frac{I_m(ql) \cdot I_{-m-1}(qx) - I_{-m}(ql) \cdot I_{m+1}(qx)}{I_m(ql) \cdot I_{-m-1}(ql) - I_m(ql) \cdot I_{m+1}(ql)} \right\} \cdot \left(\frac{x}{l}\right)^{m+1} \quad E \sin pt$$

Similarly, for the current relation we also have

$$i = \left\{ \frac{I_{-m}(ql) \cdot I_m(qx) - I_m(ql) \cdot I_{-m}(qx)}{\text{same denominator}} \right\} \frac{\beta}{(xl)^m} \cdot \frac{p}{qi} \quad E \cos pt$$

The denominator occurring in both the v and i functions can be very much simplified by means of the relation respecting conjugate functions given in Heaviside, "Electromagnetic Theory," volume 2, page 245. A text dealing with the "Symbolic Methods of Oliver Heaviside" is in course of preparation. Moreover, by introducing the P_{-m} functions deduced from the I_m functions by dividing out with the first term, the following simplified formulas result:

$$v = \left\{ P_m(ql) \cdot P_{-m}(qx) - \left(\frac{\pi x}{\lambda}\right)^2 \left(\frac{x}{l}\right)^m \cdot \right.$$

$$\left. \frac{P_{-m}(ql) \cdot P_{m+1}(qx)}{m(m+1)} \right\} \cdot E \sin pt$$

$$i = \left\{ P_m(ql) \cdot P_{-m}(qx) \left(\frac{l}{x}\right)^{2m} - P_{-m}(ql) \cdot \right.$$

$$\left. P_m(qx) \right\} \frac{\beta p}{2 m l^{2m}} \cdot E \cos pt$$

The last two formulas are especially calculable when the following general formula for P_m is used:

$$P_m(qx) = 1 + \frac{X}{1(m+1)} \left(1 + \frac{X}{2(m+2)} \left(1 + \frac{X}{3(m+3)} \left(1 + \cdots \right.\right.\right.$$

where in our case

$$X^{\frac{1}{2}} = \frac{\pi x}{\lambda} j$$

The P_m functions, Figure 3, for $m = -0.435$ have been plotted. It should be noted that if for simplicity we take

$$\frac{\pi l}{\lambda} = 1 \quad \text{then} \quad \frac{x}{l} = \frac{\pi x}{\lambda}$$

and therefore only a single argument need be used for investigational purposes. The appended curves for v and i as functions of $\frac{x}{l}$ have been plotted by means of the above formulas.

It will be noticed from the graph that if we consider 80 per cent of the length of the antenna height as measuring $\frac{1}{4}$ of the wave length, then, since $\pi l = \lambda$,

$$4 \times 0.8l = 3.2l \quad \lambda.$$

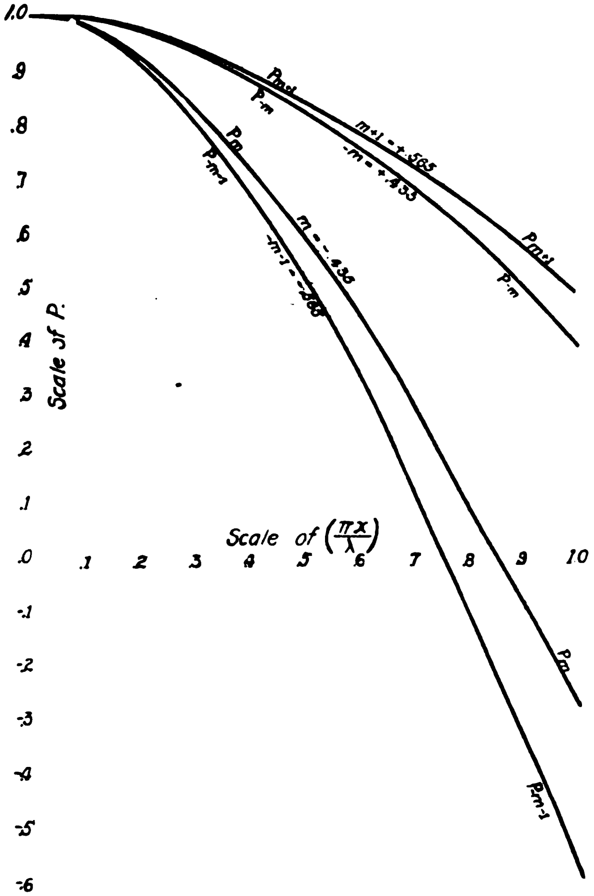

FIGURE 3

Thus, on this basis, the ordinary four times rule substantially holds, but reference should be had to measurements from the true nodal or resonant point along the antenna.

Berkeley, California,
May 29, 1919.

SUMMARY: The differential equations for an antenna of negligible resistance and leakage conductance are set up. Proceeding from previously deduced values of the inductance and capacity per centimeter along the antenna, the solution is obtained in terms of Bessel's Functions for the voltage and current distribution along the antenna. After simplification, the solution is interpreted, and curves enabling its ready employment are given.

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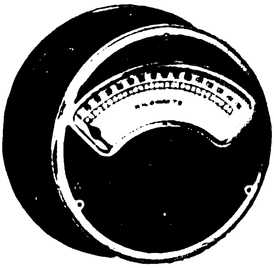

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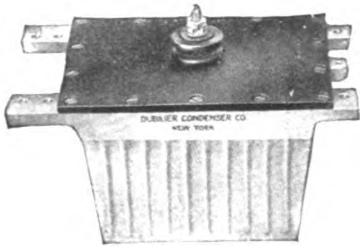

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The Institute of Radio Engineers

Volume 8 DECEMBER, 1920 Number 6

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PUBLISHED BY

THE INSTITUTE OF RADIO ENGINEERS, INC.

THE COLLEGE OF THE CITY OF NEW YORK

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MULTIPLEX RADIO TELEGRAPHY AND TELEPHONY*

By

Francis M. Ryan, J. R. Tolmie, and Roy O. Bach

(Formerly at the University of Washington, Seattle, Washington)

Various methods have been described for the simultaneous reception with a single antenna of radio signals from more than one transmitter. Similar methods have been recorded for multiplex transmission.† These methods of multiplexing may be classified as follows:

(1) Those employing several radio frequencies and obtaining selectivity by means of radio frequency tuning.

(2) Methods using a single radio frequency and several tone frequencies, selectivity being obtained by audio frequency tuning.

(3) The use of a single radio frequency modulated by several intermediate frequencies each of which has been modulated by an audio frequency. Selectivity is obtained in this case by tuning at the intermediate frequencies.

The principal example of (1) is the divided antenna method employed by Marconi for both transmission and reception and which may be used for either telegraphy or telephony. Method (2) has been used by the Telefunken Company,[1] but has never proven satisfactory owing to difficulties arising from strays. Obviously this method is not applicable to telephony. The third method was originated by Mr. R. A. Heising[2] and is applicable to both telegraphy and telephony. Altho somewhat more complicated than method (1), it is superior in so far as secrecy of transmission is concerned. A consideration of the effect of modulation shows that this method utilizes a considerable range of radio frequencies.

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* Received by the Editor, January 9, 1920. Presented before the Seattle Section of the Institute, October 15, 1919; presented before the Institute at New York, February 4, 1920.

† Since the preparation of this paper, it has come to the writer's attention that John Stone Stone patented in 1905 circuits similar to those described for multiplex transmission and reception.

[1] Zenneck, "Wireless Telegraphy," page 325.

[2] Craft and Colpitts, "Radio Telephony," "Proceedings of the American Institute of Electrical Engineers," March, 1919.

Let $\omega_1 = 2\pi \times$ (Carrier Frequency).
$\omega_2 = 2\pi \times$ (Modulation Frequency).
i = Antenna Current.
I_1 = Amplitude of the carrier frequency.
I_2 = Amplitude of the modulation frequency.
k = A constant depending on the degree of modulation.

Then $i = I_1 \sin \omega_1 t$

But I_1 is variable, $I_1 = I_2 \sin \omega_2 t + k$

$i = I_2 \sin \omega_1 t \sin \omega_2 t + k \sin \omega_1 t$

$$i = \frac{I_2}{2} \cos(\omega_1 - \omega_2) t - \frac{I_2}{2} \cos(\omega_1 + \omega_2) t + k \sin \omega_1 t$$

From this last equation, it is apparent that not only the carrier frequency will be present, but also the sum and difference of the carrier and modulation frequencies. Each of these frequencies will be further broadened by voice or other audio frequency modulation. The intermediate frequencies used are of the order of 30 kilo-cycles per second, and therefore the antenna must operate at radio frequencies differing by about 60 kilo-cycles per second. This is not possible at wave lengths exceeding about 400 meters without the use of some special device in the antenna circuit.

From these considerations it would seem that the most simple and practical method of multiplexing a radio system is the use of several radio frequencies, that is, method (1). In order to employ successfully a number of radio frequencies in the same antenna circuit, it is necessary that the antenna be resonant to each of the frequencies, that is, that its reactance be zero for the particular frequencies used. This may be accomplished by the use of certain types of impedance networks in series with the antenna. Figure 1 shows the simplest application of this method. The antenna circuit is given zero reactance for two frequencies by the loop $L_1 C_1$ coupled to the antenna by the common inductance L'_1. Figure 2 shows the equivalent lumped circuit. In this circuit the low frequency antenna capacity is assumed lumped at C_a. L_a is made equal to the sum of the inductances of the two coupling coils, (and load coil if one is used), plus one-third the inductance of the antenna for uniform current.[3] The capacity of the actual antenna is most satisfactorily measured by means of an impedance bridge. If then the natural wave length of the antenna is measured, the inductance may be calculated. At the transmitter, the two radio frequency sources may be loosely coupled to the antenna by means of the coils P_1 and P_2 as shown in Figure 1. The frequencies of these two sources must, of course, correspond to the two points of zero reactance for the antenna circuit. At the

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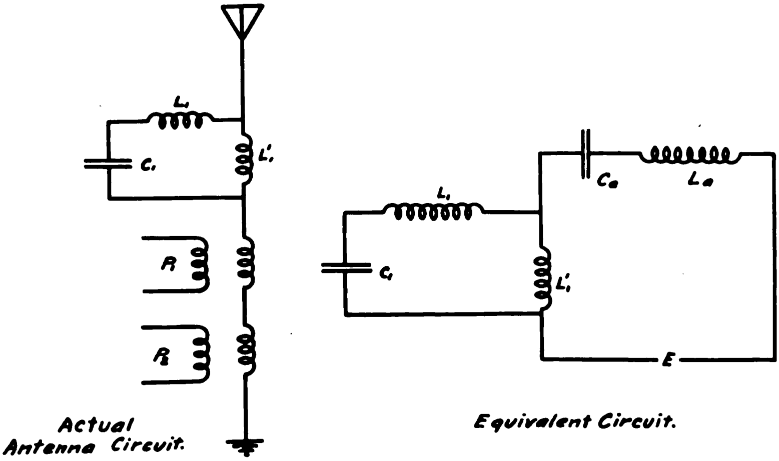
[3] Circular 74 of the Bureau of Standards, page 73.



Actual Antenna Circuit.

Equivalent Circuit.

FIGURES 1 AND 2

receiver P_1 and P_2 would be secondary coils forming parts of two closed circuits each tuned to one of the frequencies to be received. The coupling in both the transmitter and the receiver must be quite loose in order that the points of zero reactance may not be shifted or additional ones introduced.

In a circuit such as shown in Figures 1 and 2, neglecting resistance, the loop reactance is:

$$X_l = \frac{\omega L_1' - \omega^3 C_1 L_1 L_1'}{1 - \omega^2 L_1 C_1 - \omega^2 L_1' C_1}$$

The antenna reactance is:

$$X_a = \frac{\omega^2 L_a C_a - 1}{\omega C_a}$$

The total reactance of the circuit then is:

$$X_t = \frac{\omega^2 L_a C_a - 1}{\omega C_a} + \frac{\omega L_1' - \omega^3 C_1 L_1 L_1'}{1 - \omega^2 L_1 C_1 - \omega^2 L_1' C_1} \qquad (2)$$

The design of such a duplex antenna is much simplified by tuning the loop and antenna circuits independently to the same frequencies. The following familiar relations then hold:

$$K = \frac{L_1'}{\sqrt{(L_1+L_1')\,(L_a+L_1')}}$$

where K = Coupling coefficient.
L_a+L_1' = Total antenna inductance.
L_1+L_1' = Total loop inductance.

$$f_1 = \frac{f}{\sqrt{1+K}} \tag{3}$$

$$f_2 = \frac{f}{\sqrt{1-K}} \tag{4}$$

where f is the frequency to which the antenna and loop have been independently tuned, and f_1 and f_2 are the frequencies for which the reactance is zero

In order to test the practicability of such a method, a duplex radio telephone transmitter was installed in the Electrical Engineering Laboratory of the University of Washington The constants of the antenna circuit used are given in Table 1. The reactance of the equivalent antenna circuit has been calculated for various frequencies and the results are shown graphically in Figure 3. The dotted line shows the loop reactance, the dashed line the antenna reactance, and the full line the reactance of the complete circuit. Two points of zero reactance are to be noted, at 143 and 159 kilo-cycles per second, corresponding to wave lengths of 2,100 and 1,890 meters, respectively

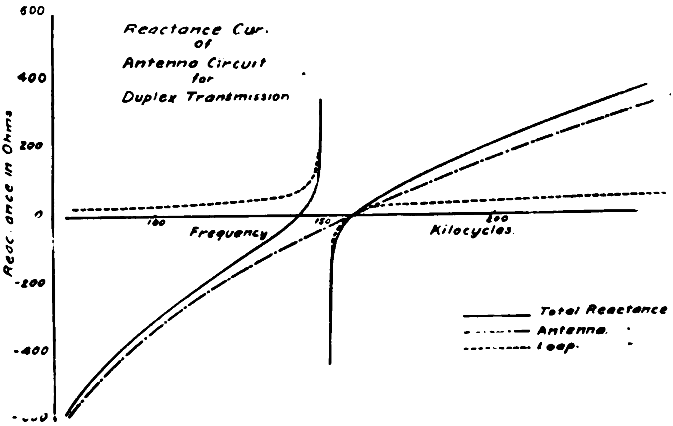


FIGURE 3

TABLE 1

$C_a = 0.00283$ microfarads
$C_1 = 0.00250$ microfarads
$L_1' = 42.1$ microhenrys
$L_a = 356.$ microhenrys
$L_1 = 408.$ microhenrys

The actual circuit employed in the transmitter is shown in Figure 4. Western Electric type "E" vacuum tubes were used, both for generation and modulation, the Heising system of modulation being employed.[4] A check of the frequency and wave length of the emitted waves was obtained by taking a resonance curve (Figure 5) with a decremeter placed in inductive relation to the transmitting antenna. The two waves as measured differ less than one per cent from the calculated values. The resonance curve is for the unmodulated waves.

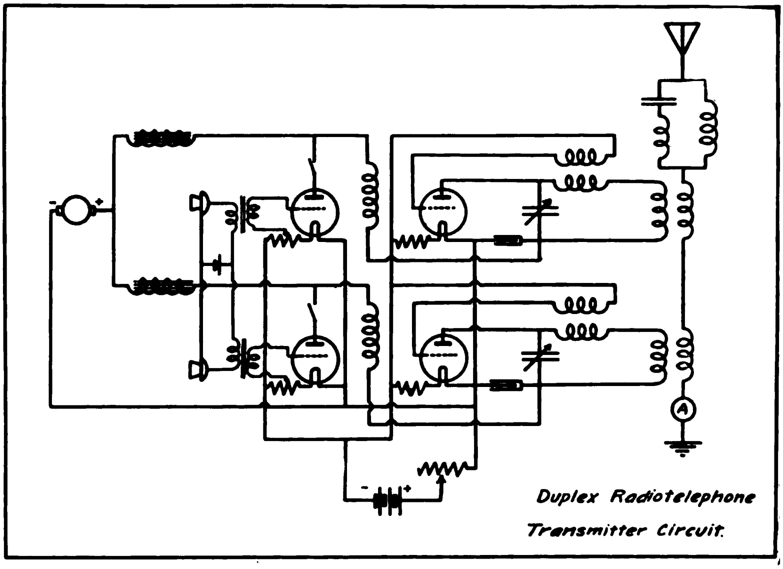


FIGURE 4

A working test was obtained by installing a duplex receiver at the Seattle Young Men's Christian Association, about five miles (8 km.) distant from the University. The circuits used are shown schematically in Figure 6. The design of the antenna circuit was similar to that at the transmitter. Two wire telephone lines were available during part of the tests, and were used for talking back to the University as no transmitting apparatus was installed at the Young Men's Christian Association.

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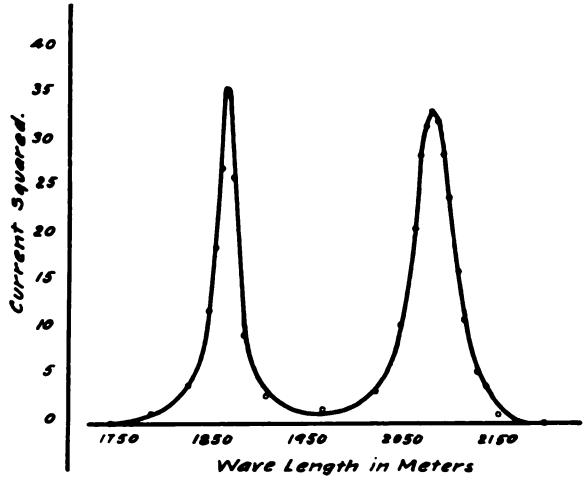
[4] Craft and Colpitts, "Radio Telephony," Proceedings of the American Institute of Electrical Engineers," March, 1919, page 360.



Resonance Curve.
of
Duplex Radio Telephone.

FIGURE 5

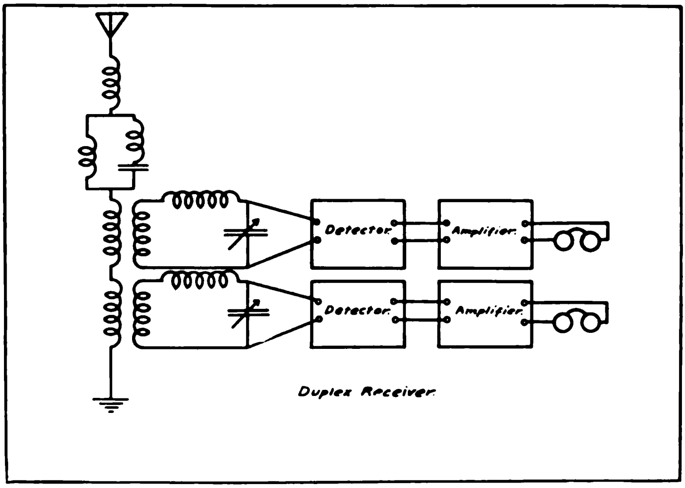


FIGURE 6

With this arrangement, two people at the Young Men's Christian Association could converse with two others at the University with practically no cross talk between the two waves used. Other tests were made allowing a phonograph to play on one of the wave lengths while the other was used for talking. Practically perfect separation was obtained. It was found that a separation of about 15 kilo-cycles between the two carrier waves was the minimum which could be satisfactorily used for telephony, but that for telegraphy the waves could be much closer together.

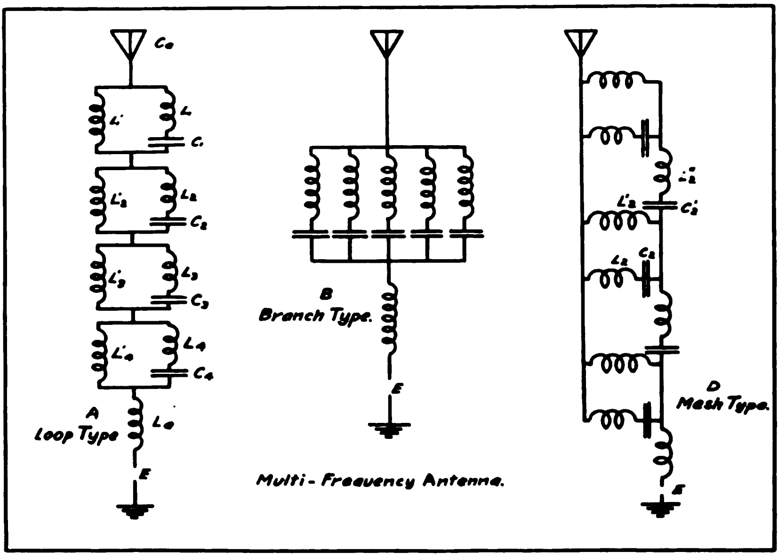


FIGURE 7

The successful completion of these tests suggested that further experiments be conducted with a greater number of frequencies. There are many networks that can be used to give the antenna the desired characteristics. Three typical ones are shown in Figure 7. The *series loop type* shown in Figure 7A gives one more point of zero reactance than there are loops in series with the antenna. The *branch type,* Figure 7B, has as many points of zero reactance as there are branches in the network. The number of points given by the *mesh type,* Figure 7D, is equal to twice the number of units, a unit being defined as consisting of a loop such as as L_2' L_2 C_2 and a series part such as L_2'' C_2' between loops.

The series loop type of network seems to offer the greatest flexibility and ease of design, and was the most thoroly investigated. An experimental network of this type used with an actual antenna to give five points of zero reactance was set up in the University laboratory. The circuit constants used are given in Table 2, the notation referring to Figure 7A. The antenna was excited by five independent vacuum tube generators inductively coupled to the antenna circuits. The coupling coils were inserted in series at the point *E* in Figure 7A. Figure 8 is a photograph of this "set up." The network is seen on the table to the left and the generator circuits and coupling coils on the table in the background. When using a single tube for each generator the operation was somewhat critical, due to reactions caused by coupling to the antenna circuit. This was overcome by using a small oscillator tube and an amplifier power tube for each of the five waves. The frequencies delivered were then independent of what kind of a system the amplifier tube fed, and depended only on the constants of the circuits of the oscillator tubes.

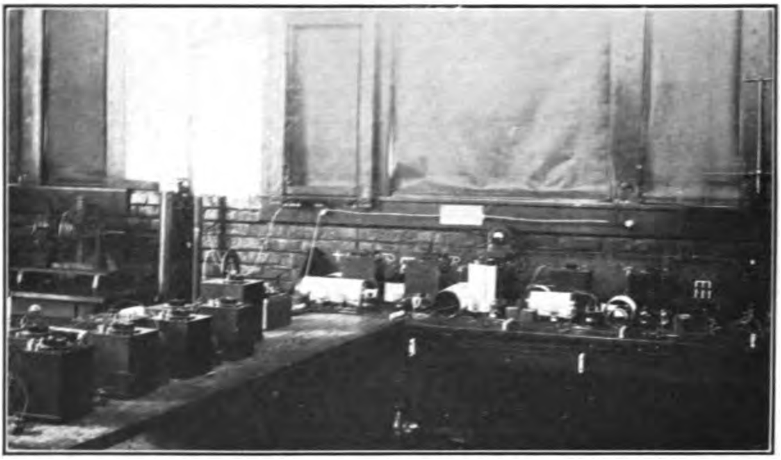


FIGURE 8—Five-wave Multiplex Radio Transmitter

TABLE 2

$C_a = 0.00127$ microfarads
$L_a = 314.$ microhenrys
$L_1 = 428.$ microhenrys
$L_2 = 522.$ microhenrys
$L_3 = 922.$ microhenrys
$L_4 = 980.$ microhenrys
$L'_1 = 155.5$ microhenrys
$L_2' = 213.$ microhenrys
$L_3' = 179.$ microhenrys
$L_4' = 157.$ microhenrys
$C_1 = 0.00148$ microfarads
$C_2 = 0.00161$ microfarads
$C_3 = 0.00141$ microfarads
$C_4 = 0.00173$ microfarads

The resonance curve shown in Figure 9 was taken by bringing a decremeter in inductive relation to the antenna. The reactance of this antenna system has been computed for a number of frequencies and the results plotted in Figures 10 and 11. Figure 10 gives the reactance curves of the separate loops and of the remainder of the circuit, and Figure 11 the reactance of the complete circuit. This last curve is obtained by adding the five curves of Figure 10 algebraically. The discontinuities or points of infinite reactance correspond to the natural frequencies of the loops and the points of zero reactance to the working points or frequencies which are to be radiated by the antenna.

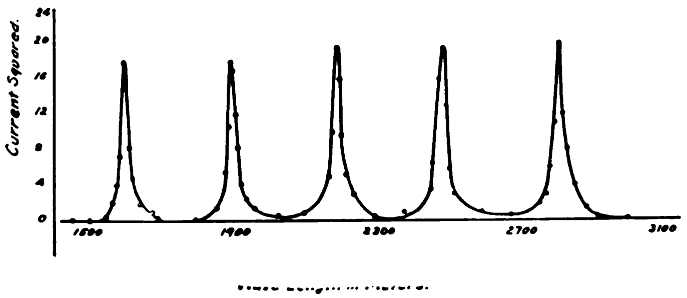


FIGURE 9

Equation (2) may be extended to apply to a network of this type having four loops by the addition of three more terms to represent the reactance of the additional loops. This equation then becomes:

$$X_t = \frac{\omega^2 L_a C_a - 1}{\omega C_a} + \frac{\omega L_1' - \omega^3 C_1 L_1 L_1'}{1 - \omega^2 C_1 L_1 - \omega^2 C_1 L_1'} + \frac{\omega L_2' - \omega^3 C_2 L_2 L_2'}{1 - \omega^2 C_2 L_2 - \omega^2 C_2 L_2'} + \frac{\omega L_3' - \omega^3 C_3 L_3 L_3'}{1 - \omega^2 C_3 L_3 - \omega^2 C_3 L_3'} + \frac{\omega L_4' - \omega^3 C_4 L_4 L_4'}{1 - \omega^2 C_4 L_4 - \omega^2 C_4 L_4'} \quad (5)$$

The notation refers to Figure 7A.

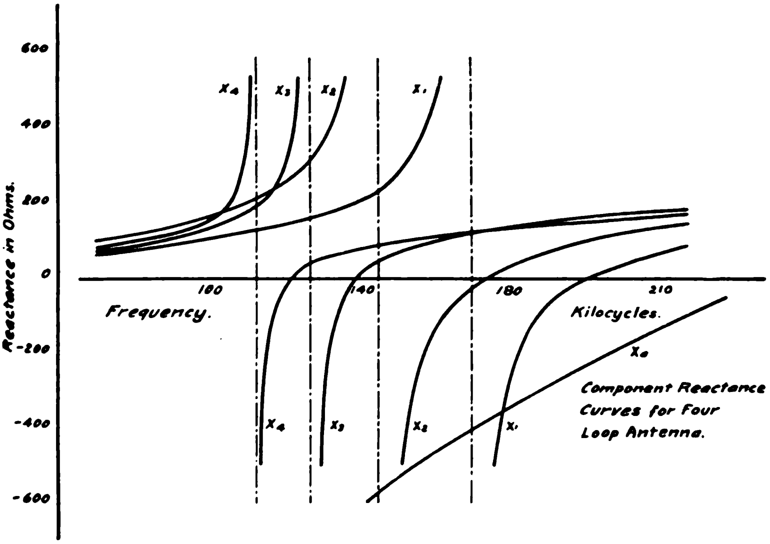


FIGURE 10

In designing a network for use with a given antenna, the problem resolves itself into one of determining the circuit constants necessary to give the points of zero reactance at the desired working frequencies or wave lengths. It will be shown

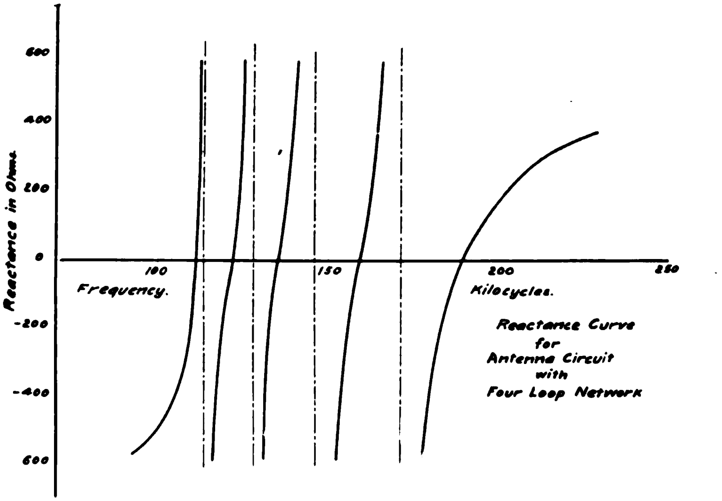


FIGURE 11

how this may be done for a system having five points of zero reactance. Reducing to a common denominator equation (5) becomes:

$$X_t = \frac{A_0\,\omega^{10} + A_2\,\omega^8 + A_4\,\omega^6 + A_6\,\omega^4 + A_8\,\omega^2 + A_{10}}{A_1\,\omega^9 + A_3\,\omega^7 + A_5\,\omega^5 + A_7\,\omega^3 + A_9\,\omega} \tag{6}$$

where:

$$A_0 = C_a\,[L_a\,\alpha_1\,\alpha_2\,\alpha_3\,\alpha_4 + L_1'\,\beta_1\,\alpha_2\,\alpha_3\,\alpha_4 + L_2'\,\alpha_1\,\beta_2\,\alpha_3\,\alpha_4$$
$$+ L_3'\,\alpha_1\,\alpha_2\,\beta_3\,\alpha_4 + L_4'\,\alpha_1\,\alpha_2\,\alpha_3\,\beta_4]$$

$$A_2 = -C_a\,[L_a\,(\alpha_1\,\alpha_2\,\alpha_3 + \alpha_1\,\alpha_2\,\alpha_4 + \alpha_1\,\alpha_3\,\alpha_4 + \alpha_2\,\alpha_3\,\alpha_4) + L_1'\,(\beta_1\,\alpha_2\,\alpha_3$$
$$+ \beta_1\,\alpha_3\,\alpha_4 + \beta_1\,\alpha_2\,\alpha_4 + \alpha_2\,\alpha_3\,\alpha_4) + L_2'\,(\beta_2\,\alpha_1\,\alpha_3 + \beta_2\,\alpha_1\,\alpha_4 + \beta_2\,\alpha_3\,\alpha_4$$
$$+ \alpha_1\,\alpha_3\,\alpha_4) + L_3'\,(\beta_3\,\alpha_1\,\alpha_2 + \beta_3\,\alpha_2\,\alpha_4 + \beta_3\,\alpha_1\,\alpha_4 + \alpha_1\,\alpha_2\,\alpha_4)$$
$$+ L_4'\,(\beta_4\,\alpha_1\,\alpha_2 + \beta_4\,\alpha_2\,\alpha_3 + \beta_4\,\alpha_1\,\alpha_3 + \alpha_1\,\alpha_2\,\alpha_3)] + \alpha_1\,\alpha_2\,\alpha_3\,\alpha_4$$

$$A_4 = C_a\,[L_a\,(\alpha_1\,\alpha_2 + \alpha_1\,\alpha_3 + \alpha_1\,\alpha_4 + \alpha_2\,\alpha_3 + \alpha_2\,\alpha_4 + \alpha_3\,\alpha_4) + L_1'\,(\beta_1\,\alpha_2$$
$$+ \beta_1\,\alpha_3 + \beta_1\,\alpha_4 + \alpha_2\,\alpha_3 + \alpha_2\,\alpha_4 + \alpha_3\,\alpha_4) + L_2'\,(\beta_2\,\alpha_1 + \beta_2\,\alpha_3$$
$$+ \beta_2\,\alpha_4 + \alpha_1\,\alpha_3 + \alpha_1\,\alpha_4 + \alpha_3\,\alpha_4) + L_3'\,(\beta_3\,\alpha_1 + \beta_3\,\alpha_2 + \beta_3\,\alpha_4 + \alpha_1\,\alpha_2$$
$$+ \alpha_1\,\alpha_4 + \alpha_2\,\alpha_4) + L_4'\,(\beta_4\,\alpha_1 + \beta_4\,\alpha_2 + \beta_4\,\alpha_3 + \alpha_1\alpha_2 + \alpha_1\,\alpha_3 + \alpha_2\,\alpha_3)]$$
$$+ \alpha_1\,\alpha_2\,\alpha_3 + \alpha_1\,\alpha_2\,\alpha_4 + \alpha_1\,\alpha_3\,\alpha_4 + \alpha_2\,\alpha_3\,\alpha_4$$

$$A_6 = -C_a\,[L_a\,(\alpha_1 + \alpha_2 + \alpha_3 + \alpha_4) + L_1'\,(\beta_1 + \alpha_2 + \alpha_3 + \alpha_4)$$
$$+ L_2'\,(\alpha_1 + \beta_2 + \alpha_3 + \alpha_4) + L_3'\,(\alpha_1 + \alpha_2 + \beta_3 + \alpha_4)$$
$$+ L_4'\,(\alpha_1 + \alpha_2 + \alpha_3 + \beta_4)] + \alpha_1\,\alpha_2 + \alpha_1\,\alpha_3 + \alpha_1\,\alpha_4 + \alpha_2\,\alpha_3$$
$$+ \alpha_2\,\alpha_4 + \alpha_3\,\alpha_4$$

$$A_8 = C_a\,[L_a + L_1' + L_2' + L_3' + L_4'] + \alpha_1 + \alpha_2 + \alpha_3 + \alpha_4$$

$$A_{10} = -1$$

$$A_1 = C_a\,\alpha_1\,\alpha_2\,\alpha_3\,\alpha_4$$

$$A_3 = -C_a\,[\alpha_1\,\alpha_2\,\alpha_3 + \alpha_1\,\alpha_2\,\alpha_4 + \alpha_2\,\alpha_3\,\alpha_4 + \alpha_1\,\alpha_3\,\alpha_4]$$

$$A_5 = C_a\,[\alpha_1\,\alpha_2 + \alpha_2\,\alpha_3 + \alpha_1\,\alpha_3 + \alpha_2\,\alpha_4 + \alpha_1\,\alpha_4 + \alpha_3\,\alpha_4]$$

$$A_7 = -C_a\,[\alpha_1 + \alpha_2 + \alpha_3 + \alpha_4]$$

$$A_9 = C_a$$

$$\alpha_1 = C_1\,L_1 + C_1\,L_1' \qquad \beta_1 = C_1\,L_1$$
$$\alpha_2 = C_2\,L_2 + C_2\,L_2' \qquad \beta_2 = C_2\,L_2$$
$$\alpha_3 = C_3\,L_3 + C_3\,L_3' \qquad \beta_3 = C_3\,L_3$$
$$\alpha_4 = C_4\,L_4 + C_4\,L_4' \qquad \beta_4 = C_4\,L_4$$

In order that the effective resistance of the antenna system may not be excessive the natural frequencies of the loops must differ considerably from the working frequencies, that is, it is desirable to place the points of infinite reactance about midway between those of zero reactance. The points of infinite reactance are determined by the roots of the denominator of equation (6) equated to zero. If R_1, R_2, R_3, and R_4 are the squares of the respective values of ω for which the reactance is desired infinite, then the following relations satisfy the condition that the denominator equals zero, thereby fixing the infinite points.

$$\begin{aligned} a_1 &= \frac{1}{R_1} \\ a_2 &= \frac{1}{R_2} \\ a_3 &= \frac{1}{R_3} \\ a_4 &= \frac{1}{R_4} \end{aligned} \tag{7}$$

This group of equations determines the values of a_1, a_2, a_3, and a_4; and when the frequencies for which the reactance is to be infinite are decided upon, the a's become numerics and their values may be substituted in the numerator.

The frequencies for which the reactance is zero are determined by the roots of the numerator of equation (6) equated to zero. If the squares of the values of ω for which the reactance is desired zero are r_1, r_2, r_3, r_4, and r_5, then the following relations, follow from the theory of equations:

$$\begin{aligned} \frac{A_2}{A_o} &= -(r_1+r_2+r_3+r_4+r_5) \\ \frac{A_4}{A_o} &= r_1 r_2+r_1 r_3+r_1 r_4+r_1 r_5+r_2 r_3+r_2 r_4+r_2 r_5+r_3 r_4+r_3 r_5+r_4 r_5 \\ \frac{A_6}{A_o} &= -(r_1 r_2 r_3+r_1 r_2 r_4+r_1 r_2 r_5+r_1 r_3 r_4+r_1 r_3 r_5+r_1 r_4 r_5 \\ &\qquad +r_2 r_3 r_4+r_2 r_3 r_5+r_2 r_4 r_5+r_3 r_4 r_5) \\ \frac{A_8}{A_o} &= r_1 r_2 r_3 r_4+r_1 r_3 r_4 r_5+r_1 r_2 r_4 r_5+r_1 r_2 r_3 r_5+r_2 r_3 r_4 r_5 \\ \frac{A_{10}}{A_o} &= -r_1 r_2 r_3 r_4 r_5 \end{aligned} \tag{8}$$

This group consists of five simultaneous equations in the unknowns β_1, β_2, β_3, β_4, L_a, L_1', L_2', L_3', L_4', and C_a. Five of these may be given arbitrary values providing care is taken not to choose values which lead to physically impossible solutions. It has been found convenient to give values to β_1, β_2, β_3, and β_4 and to solve for L_a, L_1', L_2', L_3', and L_4' in terms of C_a. The solution may then be carried out by means of determinants.

Equation (5) may be extended to apply to a general case giving $\frac{n}{2}$ points of zero reactance and having $\frac{n}{2}-1$ loops in circuit.

$$X_t = \frac{\omega^2 L_a C_a - 1}{\omega C_a} + \frac{\omega L_1' - \omega^3 L_1 C_1 L_1'}{1 - \omega^2 L_1 C_1 - \omega^2 L_1' C_1} + \cdots$$
$$+ \cdots \frac{\omega L_m' - \omega^3 L_m C_m L_m'}{1 - \omega^2 L_m C_m - \omega^2 L_m' C_m} \qquad (9)$$

Reducing to a common denominator equation (9) becomes of the following general form:

$$X_t = \frac{A_o \omega^n + A_2 \omega^{n-2} + A_4 \omega^{n-4} \cdots + A_{n-2} \omega^2 + A_n}{A_1 \omega^{n-1} + A_3 \omega^{n-3} + A_5 \omega^{n-5} \cdots + A_{n-3} \omega^3 + A_{n-1} \omega} \qquad (10)$$

The coefficients A_o, A_1, A_2, and so on, are functions of the electrical constants of the system as in the case of the four-loop network. The roots of the denominator equated to zero merely show that the points of infinite reactance occur whenever the frequency is equal to the natural frequency of one of the loops. In a manner similar to that shown for the four-loop network, equations may be set up relating the roots and the coefficients of the numerator. There will be $\frac{n}{2}$ of these equations and n variables. In general, this system of equations will be indeterminate, but by reducing the complexity of the network or by imposing certain arbitrary conditions upon it, the system will become determinate. The complexity of the network may be reduced by removing the inductance from the capacity sides of the loops. This is not usually advantageous as it often leads to inconvenient magnitudes for the remaining circuit constants. Arbitrary values may be assigned (within limits) to β_1, β_2, β_3, and so on, as in the four-loop case, and then the values of L_a, L_1', L_2', and so on, may be determined in terms of C_a. The solution is then applicable to any suitable antenna. The values of L_1', L_2', L_3', and so on, can be assigned and solution made for L_a, β_1, β_2, β_3, and so on, if desired.

Equation (10) can be written in the following condensed form.

$$X_t = \frac{\sum_{q=0}^{n/2} A_{2q} \omega^{n-2q}}{\sum_{q=2p+1}^{n-1} A_q \omega^{n-q}} \qquad (11)$$

The equivalent expansion for the branch type network will be:

$$X_t = \frac{\sum\limits_{q=2p+1}^{n-1} A_q \, \omega^{n-q+2}}{\sum\limits_{q=0}^{n} A_{2q} \, \omega^{n-2q}} \qquad (12)$$

Equation (12) differs from the reciprocal of equation (11) only in the degree of the odd terms. If, however, the series loop type of circuit, Figure 7A, contains capacity in both branches of the loop its reactance is given by equation (12) instead of equation (11). The mesh type of circuit shown in Figure 7D gives a reactance expansion similar to equation (11).

The following numerical solution, of an antenna with a two-loop network is given to illustrate the method of design and calculation of multi-frequency circuits. The results were checked experimentally to show the degree of accuracy that might be expected.

The desired working frequencies were chosen to correspond to wave lengths of 1,600, 1,900, and 2,200 meters. The frequencies for infinite reactance, that is, the frequencies of the respective loops, were taken to correspond to wave lengths of 1,750 and 2,050 meters. β_1 and β_2, that is, the product of inductance and capacity on the condenser side of the loops, were selected for wave lengths of 1,550 and 1,850 meters.

It is first necessary to obtain the equation for the total reactance of the entire network. This is as follows:

$$X_t = \frac{\omega^2 L_a C_a - 1}{\omega C_a} + \frac{\omega L_1' - \omega^3 L_1 C_1 L_1'}{1 - \omega^2 L_1 C_1 - \omega^2 L_1' C_1} + \frac{\omega L_2' - \omega^3 L_2 C_2 L_2'}{1 - \omega^2 L_2 C_2 - \omega^2 L_2' C_2} \qquad (13)$$

In a manner exactly similar to that previously shown for the four-loop network three simultaneous equations are obtained from equation (13).

$$L_a \left[a_1 + a_2 + \frac{A_2}{A_o} a_1 a_2 \right] + L_1' \left[a_2 + \beta_1 + \frac{A_2}{A_o} a_2 \beta_1 \right] + L_2' \left[a_1 + \beta_2 + \frac{A_2}{A_o} a_1 \beta_2 \right] = - \frac{a_1 a_2}{C_a}$$

$$L_a \left[1 - \frac{A_4}{A_o} a_1 a_2 \right] + L_1' \left[1 - \frac{A_4}{A_o} a_2 \beta_1 \right] + L_2' \left[1 - \frac{A_4}{A_o} a_1 \beta_2 \right] = - \frac{a_1 + a_2}{C_a}$$

$$L_a a_1 a_2 \frac{A_6}{A_o} + L_1' a_2 \beta_1 \frac{A_6}{A_o} + L_2' a_1 \beta_2 \frac{A_6}{A_o} = - \frac{1}{C_a}$$

These equations are of the following form:

$$L_a a + L_1' b + L_2' C = \frac{k}{C_a}$$

where

$a_1 = -5.21983 \times 10^{-12}$ $b_1 = -4.35003 \times 10^{-12}$ $c_1 = -4.41122 \times 10^{-12}$
$a_2 = -2.17443$ $b_2 = -1.49029$ $c_2 = -1.58529$
$a_3 = -1.02424 \times 10^{12}$ $b_3 = -0.80364 \times 10^{12}$ $c_3 = -0.8343 \times 10^{12}$

$k_1 = -1.02155 \times 10^{-24}$
$k_2 = -2.04676 \times 10^{-12}$
$k_3 = -1$

Solving these equations by determinant for L_a, L_1', and L_2', we obtain

$$L_a = \frac{0.00884 \times 10^{-12}}{C_a \times 0.01227} \text{ henrys}$$

$$L_1' = \frac{0.002398 \times 10^{-12}}{C_a \times 0.01227} \text{ henrys}$$

$$L_2' = \frac{0.00264 \times 10^{-12}}{C_a \times 0.01227} \text{ henrys}$$

For $C_a = 0.00283$ microfarads
$L_a = 255$ microhenrys
$L_1' = 69$ microhenrys
$L_2' = 76$ microhenrys

The values of L_1, L_2, C_1, and C_2 are then found from the known values of α_1, α_2, β_1, and β_2.

TABLE 3

λ (Assumed)	λ (Experimental)	Per Cent. Error
1,600	1,620	1.2
1,900	1,900	0.0
2,200	2,260	2.7

The discrepancies observed in the experimental check are probably due to the effect of long leads and to the fact that some of the constants were measured at audio frequencies.

The previous equations may be considerably simplified by making $\beta_1 = \beta_2 = L_a C_a$. A network of this type was set up and checked experimentally. The results obtained are as follows:

TABLE 4

λ (Assumed)	λ (Experimental)	Per Cent. Error
1,600	1,605	0.3
1,900	1,905	0.2
2,200	2,208	0.4

It is believed that such systems of multiplexing by means of multi-frequency antennas will be of considerable value for fleet communication in the Navy. It should also prove valuable in marine coast stations designed to connect ship radio telephone stations with the existing land wire telephone system. Another possible application is that of trans-oceanic communication at long wave lengths.

In high power stations, the output is often limited by the corona voltage of the antenna. It is to be noted that in using a multi-frequency antenna for multiplex transmission, the maximum instantaneous voltage will be the arithmetic sum of the maximums of the several frequencies impressed. This will occur at a relatively low frequency, but demands consideration. Another limitation of the methods proposed is the lack of flexibility of adjustment. The effective resistance of such networks becomes excessive when it is attempted to place the working waves very close together. Some additional work should be done studying the effective resistance and efficiency of multi-frequency antennas. In spite of these limitations the method will probably find considerable application in special instances.

The writers are deeply indebted to Dr. C. E. Magnusson and Prof. L. F. Curtis of the Electrical Engineering Department of the University of Washington for the excellent facilities provided for the experiments described, and for their interest in the work. Thanks is also due to Mr. V. I. Kraft, of the Seattle Young Men's Christian Association for his generosity in allowing the use of their antenna and premises. It should also be mentioned that the Navy Department, thru Lieutenant-Commander Frank Luckel, District Communication Superintendent at the Navy Yard, Puget Sound, very kindly extended the privilege of using the transmitter. The Office of the Chief Signal Officer of the Army co-operated by furnishing the necessary vacuum tubes without which the experiments would have been impossible.

University of Washington,
Seattle, Washington,
December 16, 1919

SUMMARY: The authors classify the chief methods of multiplex radio communication as multi-radio frequency, mono-radio-frequency with multi-audio-frequency, and mono-radio frequency modulated at several super-audio-frequencies each of which is itself modulated at audio frequency.

The authors classify antenna systems for the first method as of the series loop type, the branch type, and the mesh type. Each of these systems has zero reactance at a number of radio frequencies. The series type is investigated mathematically for several cases and checked experimentally with good accuracy.

Duplex and quintuplex radio telephony and telegraphy over 5 miles (8 km.) at wave lengths around 2,000 meters were experimentally accomplished, and are described.

A CONTRIBUTION TO THE THEORY OF MAGNETIC FREQUENCY CHANGERS *

By

J. Zenneck

(Professor of Experimental Physics, Institute of Technology, Munich, Germany)

I. ARRANGEMENT OF CIRCUITS

The arrangement on which the following theoretical considerations are based, is well known.† It consists of two iron cores, *A* and *B* (Figure 1), on each of which are wound three coils; a primary coil (N_1), a direct current coil (N_0) and a secondary coil (N_2).

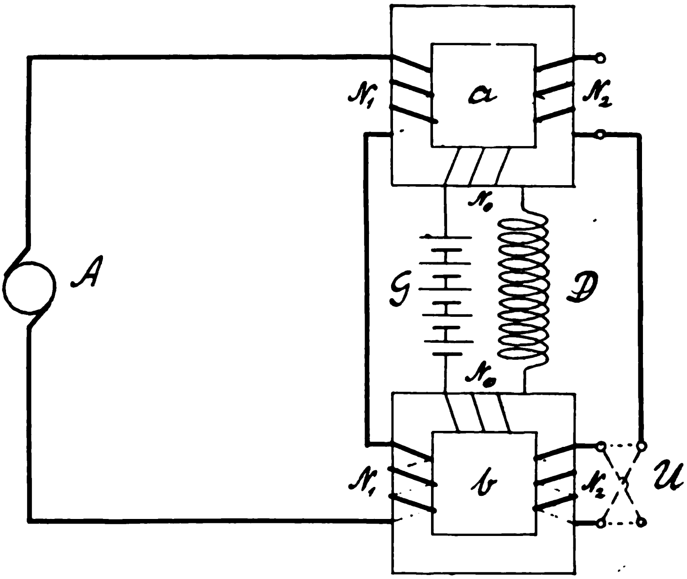


Figure 1

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* Presented before The Institute of Radio Engineers, New York, September 1, 1915.

†See "Radio Frequency Changers," by A. N. Goldsmith, Proceedings of The Institute of Radio Engineers, Volume 3, page 55.

The two primary coils N_1 are connected in series, and inserted in the circuit of the radio-frequency alternator A. The direct current coils N_0 are fed by a direct current source G, the circuit of which contains a choke coil D. These coils are wound on the two iron cores in such a way that if, at a given moment the direct current in the iron core A assists the magnetic field of the primary current, in the iron core B the magnetic fields of the direct current and that of the primary current are opposed to each other. The secondary coils N_2 can either be connected in series (assisting) or in opposition by a switch U.

II. ASSUMPTIONS

1. The inductance of the coil D is assumed to be so large that no radio-frequency current of any appreciable amplitude is allowed to flow in the direct current circuit.

2. The size and material of the two iron cores are alike and the corresponding coils have the same number of turns (N_1, N_o, N_2).

3. The magnetic field intensity in the iron cores being H, the magnetic induction B cannot be assumed as being proportional to H. In fact, the operation of the magnetic frequency changers is due precisely to the fact that such proportionality does not exist. The relation between the magnetic induction and the magnetic field intensity may therefore be expressed by an equation:

$$B = sH - s'H^3 \quad . \quad . \quad . \quad . \quad . \qquad (1)$$

where s and s' are constants of the iron used.

The magnetic flux in the iron cores (Φ_a and Φ_b) is proportional to B, and the magnetic field intensity H is proportional to the number of ampere turns (F_a, F_b). We may therefore write

$$\left.\begin{array}{l}\Phi_a = pF_a - qF_a^3 \\ \Phi_b = pF_b - qF_b^3\end{array}\right\} \quad . \quad . \quad . \quad . \quad . \qquad (2)$$

p and q being constants.

The curve corresponding to this equation has a maximum for

$$\left.\begin{array}{l}F_a \\ F_b\end{array}\right\} = \sqrt{\frac{p}{3q}} = S$$

Evidently we are not entitled to use equation (2) beyond this point, as then by a further increase of the ampere turns, a decrease of the magnetic flux would be produced. This implies that always

$$\left.\begin{array}{l}F_a \\ F_b\end{array}\right\} \leqq S \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (3)$$

Substituting the maximum value S in equation (2), we get

$$\left.\begin{aligned}\Phi_a &= pF_a\left[1-\frac{1}{3}\left(\frac{F_a}{S}\right)^2\right]\\ \Phi_b &= pF_b\left[1-\frac{1}{3}\left(\frac{F_b}{S}\right)^2\right]\end{aligned}\right\} \quad \ldots\ldots \quad (4)$$

The curve corresponding to this equation is shown in Figure 2.

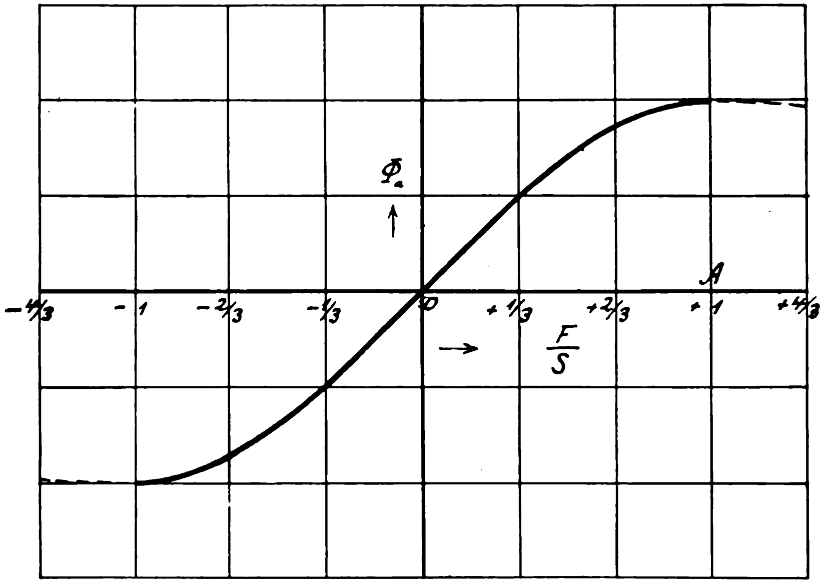

FIGURE 2

III. EMF. IN THE SECONDARY COILS

As the secondary coils in Figure 1 are not inserted in a closed circuit, no secondary current i_2 of any appreciable amplitude is present. The ampere turns F_a and F_b are therefore only due to the primary current i_1 and the direct current I_o, that is

$$\left.\begin{aligned}F_a &= N_1 i_1 + N_o I_o\\ F_b &= N_1 i_1 - N_o I_o.\end{aligned}\right\} \quad \ldots\ldots\ldots \quad (5)$$

As to the primary current, the assumption may be made that it is of substantially sinusoidal form; that is

$$i_1 = I_1 \sin \omega t \quad \ldots\ldots \quad (6)$$

This assumption seems to be harmless, but is not so under the conditions of equation (4). It is easy to show that under these conditions even when the voltage of the alternator is exactly sinusoidal, in the emf. of the primary circuit a third harmonic is produced, the amplitude of which is of the same order of magnitude as that of the voltage of the alternator. There is indeed but one means for substantially realising the condition of equation (6): namely, inserting into the primary circuit a condenser C_1 and an inductance L_1, thereby substituting the arrangement of Figure 3 for that of Figure 1, and adjusting the capacity C_1 and the inductance L_1 so that the primary circuit is at least approximately in resonance with the frequency of the alternator. The use of equation (6) is therefore restricted to this case.*

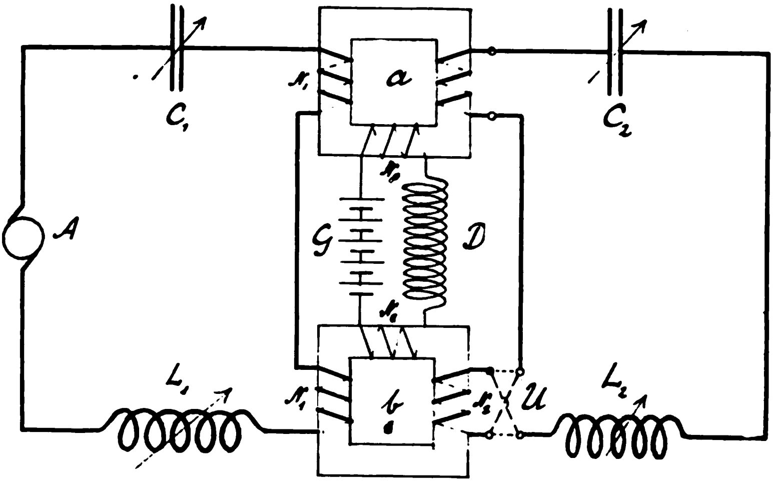

FIGURE 3

a. Connecting first the secondary coils *in series, and additively*, the emf. e_2 induced in them, is determined by the equation

$$e_2 = -N_2 \frac{d}{dt}(\Phi_a + \Phi_b)$$

Now by equations (4), (5) and (6)

$$\Phi_a = p\,(N_1 i_1 + N_o I_o)\left\{1 - \frac{1}{3}\left(\frac{N_1 i_1 + N_o I_o}{S}\right)^2\right\}$$

$$\Phi_b = p\,(N_1 i_1 - N_o I_o)\left\{1 - \frac{1}{3}\left(\frac{N_1 i_1 - N_o I_o}{S}\right)^2\right\}$$

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* I am well aware that the mathematical method of this paper is substantially correct only as long as the amplitudes of the current harmonics are small compared to that of the fundamental.

and therefore

$$\Phi_a + \Phi_b = 2\,p\,N_1\,i_1 \left\{ 1 - \left(\frac{N_a I_a}{S} \right)^2 - \frac{1}{3} \left(\frac{N_1 i_1}{S} \right)^2 \right\}$$

$$= 2\,p\,N_1\,I_1 \sin \omega t \left\{ 1 - \left(\frac{N_a I_a}{S} \right)^2 - \frac{1}{3} \left(\frac{N_1 I_1 \sin \omega t}{S} \right)^2 \right\}$$

Since

$$\sin^3 \omega t = \frac{3}{4} \sin \omega t - \frac{1}{4} \sin 3\,\omega t,$$

we get

$$\Phi_a + \Phi_b = 2\,p\,N_1\,I_1 \left\{ 1 - \left(\frac{N_a I_a}{S} \right)^2 - \frac{1}{4} \left(\frac{N_1 I_1}{S} \right)^2 \right\} \sin \omega t$$

$$- \frac{p}{6} S \left(\frac{N_1 I_1}{S} \right)^3 \sin 3\,\omega t$$

and therefore

$$e_2 = -2\,p\,\omega\,N_1\,N_2\,I_1 \left\{ 1 - \left(\frac{N_a I_a}{S} \right)^2 - \frac{1}{4} \left(\frac{N_1 I_1}{S} \right)^2 \right\} \cos \omega t$$

$$+ \frac{1}{2}\,p\,\omega\,N_2 S \left(\frac{N_1 I_1}{S} \right)^3 \cos 3\,\omega t.$$

This means that in addition to an emf. of the fundamental frequency of the alternator an emf. of a frequency three times greater than that of the alternator is produced. The arrangement discussed is therefore a device for *tripling* the frequency.

The amplitude of the emf. of triple frequency is proportional to the third power of the primary amplitude, but does not depend on the direct current. If therefore no direct current at all were used, the same amplitude of the emf. of triple frequency would result. The only function of the direct current is to diminish the amplitude of the unwelcome emf. of the fundamental frequency; but as long as the restriction expressed by equation (3) holds, this amplitude cannot be reduced to zero.

b. Connecting the secondary coils *in opposition*, the emf. e_2 induced in them is determined by

$$e_2 = -N_2 \frac{d}{d\,t} (\Phi_a - \Phi_b).$$

As

$$\Phi_a - \Phi_b = 2\,p\,N_o I_o \left\{ 1 - \left(\frac{N_1 I_1}{S} \right)^2 - \frac{1}{3} \left(\frac{N_o I_o}{S} \right)^2 \right\}$$

$$= 2\,p\,N_o I_o \left\{ 1 - \left(\frac{N_1 I_1 \sin \omega t}{S} \right)^2 - \frac{1}{3} \left(\frac{N_o I_o}{S} \right)^2 \right\}$$

$$= 2\,p\,N_o I_o \left\{ 1 - \frac{1}{2} \left(\frac{N_1 I_1}{S} \right)^2 + \frac{1}{2} \left(\frac{N_1 I_1}{S} \right)^2 \cos 2\,\omega t - \frac{1}{3} \left(\frac{N_o I_o}{S} \right)^2 \right\},$$

we get

$$e_2 = 2\,p\,\omega\,N_2 S \left(\frac{N_o I_o}{S} \right) \left(\frac{N_1 I_1}{S} \right)^2 \sin 2\,\omega t \quad \cdot \cdot \quad \cdot \cdot \qquad (7)$$

This equation shows first, that as long as the primary current is purely sinusoidal, the emf. induced in the secondary coils is purely sinusoidal and of twice the frequency of the primary current. The arrangement represented is therefore a device for *doubling* the frequency and an ideal one; as neither the fundamental frequency nor harmonics higher than the second are present in the emf. induced in the secondary coils.

Further, it is evident from equation (7) that for the production of this emf., the direct current is absolutely necessary. For by making the direct current zero, the amplitude of e_2 also would be reduced to zero.

Finally, the amplitude of e_2 being proportional to I_o and to I_1^2, and I_o and I_1 being limited by the condition (3), that is

$$N_1 I_1 + N_o I_o \leqq S \qquad \text{or}$$

$$\frac{N_1 I_1}{S} + \frac{N_o I_o}{S} \leqq 1,$$

The question arises, which ratio of $\frac{N_1 I_1}{S}$ and $\frac{N_o I_o}{S}$ produces the maximum amplitude of e_2. This question is answered by Figure 4, in which the abscissas are proportional to $\frac{N_1 I_1}{S}$, the ordinates proportional to E_2, the amplitude of e_2, and the assumption has been made, that the value of $\frac{N_1 I_1}{S} + \frac{N_o I_o}{S}$ is the highest possible, namely 1. The curve in Figure 4 shows that the amplitude E_2 reaches a marked maximum for $\frac{N_1 I_1}{S} = \frac{2}{3}$, that is, (compare Figure 2), when the magnetic flux Φ_a due to the alternating primary current varies from the value corresponding to $F_a = +S$ to the value corresponding to $F_a = -\frac{1}{3}S$ and Φ_b from $+\frac{1}{3}S$ to $-S$.

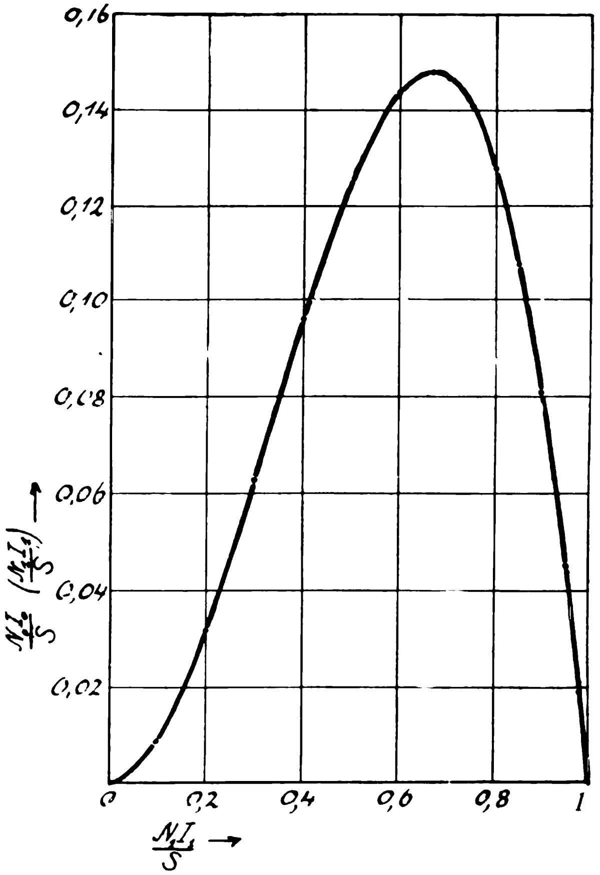

FIGURE 4

IV. THE PRIMARY CIRCUIT

In calculating the primary current i_1 as a function of the emf. e_1 of the alternator, the secondary circuit being disconnected, we restrict ourselves to the case, where the primary circuit is at least approximately in resonance with the frequency of the alternator. In this case we may neglect all harmonics of the current. We retain the equation

$$i_1 = I_1 \sin \omega t$$

and assume $\qquad e_1 = E_1 \sin(\omega t + \phi).$

Then the amplitude I_1 and the phase difference ϕ are the terms to be calculated.

a. The differential equation for the condenser circuit of Figure 3, acted upon by an emf. e_1 is

$$R_1 \frac{d\,i_1}{d\,t} + L_1 \frac{d^2\,i_1}{d\,t^2} + N_1 \frac{d^2}{d\,t^2}(\Phi_a + \Phi_b) + \frac{i_1}{C_1} = \frac{d\,e_1}{d\,t} \cdot \cdot \cdot \cdot \cdot \quad (8)^*$$

Under the assumptions just mentioned, and using for $(\Phi_a + \Phi_b)$ the term calculated in *IIIa*, its solution is

$$I_1 = \frac{E_1}{\sqrt{R_1^2 + \left[\omega\left(L_1 + L_1'\left\{1 - \left(\frac{N_o I_o}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2\right\}\right) - \frac{1}{\omega C_1}\right]^2}}$$

$$\tan\phi = \frac{\omega\left(L_1 + L_1'\left\{1 - \left(\frac{N_o I_o}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2\right\}\right) - \frac{1}{\omega C_1}}{R_1} \qquad (9)$$

where $L_1' = 2\,p\,N_1^2$. The term

$$L_1'\left\{1 - \left(\frac{N_o I_o}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2\right\}$$

is to be considered as being the inductance due to the iron cores.

If no direct current were used, and if the amplitude of the primary current were so small that $\frac{N_1 I_1}{S} \ll 1$, we would have obtained the ordinary equations for a condenser circuit with the constant inductance $L_1 + L_1'$, i.e.:

$$I_1 = \frac{E_1}{\sqrt{R_1^2 + \left[\omega(L_1 + L_1') - \frac{1}{\omega C_1}\right]^2}}$$

$$\tan\phi = \frac{\omega\,(L_1 + L_1') - \frac{1}{\omega C_1}}{R_1} \qquad (10)$$

Comparing equations (9) and (10), it is easily seen that the direct current has not altered conditions greatly; it only diminishes the inductance due to the iron cores by the constant term $-\left(\frac{N_o I_o}{S}\right)^2$. But the term caused by the primary current, $-\frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2$, changes the whole situation as soon as an appreciable primary current is present and therefore $\frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2$ is no longer $\ll 1$.

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* L_1 being the inductance of the primary circuit outside of the frequency changer.

b. One difference is immediately seen, when we specify the resonance condition; that is, the condition which makes the ratio $I_1 : E_1$ a maximum.

In the ordinary case of constant inductance (equation 10) this condition is:

$$\omega(L_1+L_1') - \frac{1}{\omega C_1} = 0, \quad \cdot \quad \cdot \quad \cdot \quad \cdot \qquad (11)$$

but in our case it is represented by

$$\omega\left(L_1+L_1'\left\{1-\left(\frac{N_0 I_0}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2\right\}\right) - \frac{1}{\omega C_1} = 0 \cdot \quad \cdot \quad \cdot \qquad (12)$$

While in the ordinary case of constant inductance, there is one definite resonance frequency independent of the E.M.F. of the alternator, in our case, there is no general resonance condition at all. The resonance frequency, as might have been expected before hand, depends on the amplitude of the primary current and therefore on the excitation of the alternator. When for one voltage of the alternator the circuit has been adjusted so as to be in resonance to the frequency of the alternator, it is no longer in resonance as soon as the excitation, and therefore the voltage of the alternator, is changed.

In one respect, our case and the ordinary one are alike. When the resonance condition is fulfilled, the amplitude of the current is determined only by the effective resistance R_1 and the equations (9) as well as (10) become

$$I_1 = \frac{E_1}{R_1}; \quad \tan\phi = 0.$$

c. Nevertheless, owing to the fact that because of the term $-\frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2$, the inductance is no longer constant, the whole situation is changed in a very fundamental way. This can be best shown by drawing for different frequencies the characteristics of the circuit; that is, by plotting curves having as abscissas the amplitudes (or R.M.S. values) I_1 of the current and as ordinates the amplitudes (or R.M.S. values) E_1 of the E.M.F. of the alternator.

In the case of constant inductance, according to equation (10), all these curves would be straight lines, and all points fulfilling the resonance condition (11) would be situated on one of them; namely on the straight line represented by the equation $I_1 = \frac{E_1}{R_1}$.

In our case, corresponding to equation (9) the state of affairs is quite different, as is seen from Figure 5, which has been calculated under the following assumptions:

$$R_1 = 5\,\text{ohms};\ C_1 = 0.1\ \mu\text{f. (mfd.)};\ \frac{L_1'}{L_1} = 10;$$

$$\frac{N_0 I_0}{S} = \frac{1}{3};\ \frac{N_1 I_{1(max)}}{S} = \frac{2}{3}\ \text{for}\ I_{1(max)} = 100\ \text{amperes};$$

$$\omega\left(L_1 + L_1'\left\{1 - \left(\frac{N_0 I_0}{S}\right)^2\right\}\right) - \frac{1}{\omega C_1} = 0$$

for $\omega = 2\pi \times 10{,}000$ cycles per sec.

This means that the primary current being so small that $\frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2 \ll 1$, the circuit is in resonance to the frequency 10,000 cycles per second.

All the characteristic curves of Figure 5 are curved lines, many of them with a falling portion. A falling characteristic of a circuit means unstable conditions.* All points, therefore, which are situated on a falling part of such a curve, altho they are actually given by equation (9), cannot be realized in practice, at least so far as stationary conditions are concerned.

It will be noted that the resonance points defined by equation (12) are nothing more than the points, where the straight line corresponding to the equation

$$I_1 = \frac{E_1}{R_1} = \frac{E_1}{5}$$

(that is, the dashed line *OA* in Figure 5) is a tangent to the characteristic curves. In those curves in which these points are very near to the boundary between the rising and the falling portions, the resonance is theoretically stable, as well as in the curves which are entirely rising. Yet from a practical point of view, such resonance may be considered as unstable, since a very small variation of the current or the voltage may cause instability, and therefore no operator would be willing to work the set under such delicate conditions of adjustment.

The curves of Figure 5 show another feature, which may be interesting from a practical point of view. Suppose the normal voltage of the alternator to be about 500 volts. Then the most favorable operating conditions would be those represented by the point A of the curve $n = 10{,}600$; as at this point the maximum current of 100 amperes would be produced by the normal voltage. But this point is so near to the falling, and therefore unstable part of the curve, that no operator would take the risk of working at this point. Therefore, from a practical point of

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* In the following discussion, for the sake of simplicity, the characteristic of the alternator is assumed to be a straight line parallel to the I axis.

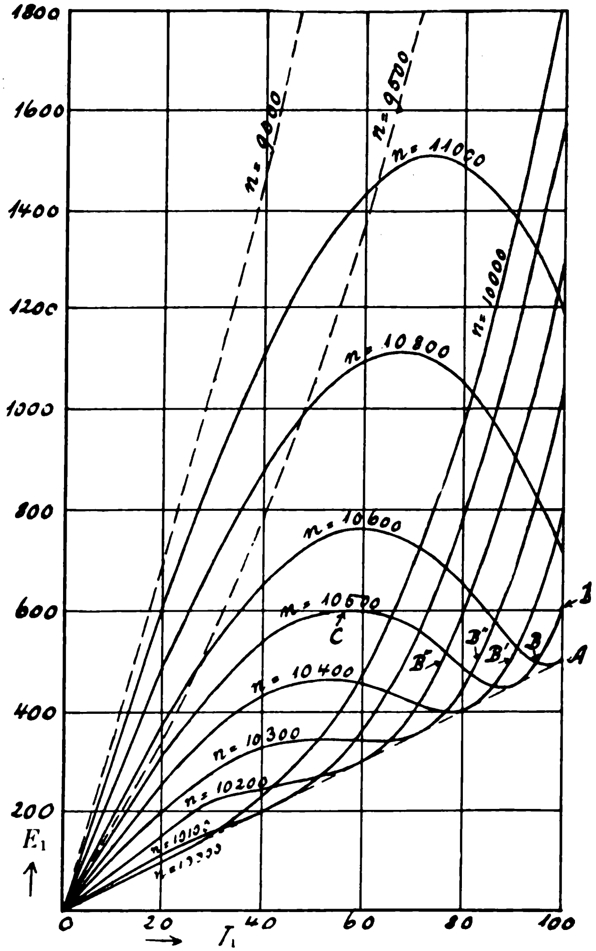

FIGURE 5

view, the best point on the curves is the point B on the curve $n = 10{,}500$. To reach this point, we may either over-excite the alternator up to 600 volts while keeping its frequency constant at 10,500 cycles per second and then proceed along the curve

$n_1 = 10{,}500$ up to the point C. Then, as soon as the voltage is further increased slightly, the current will jump from the value of about 56 amperes at the point C to the value of about 100 amperes at the point D. By then decreasing the excitation of the alternator to the normal value required for 500 volts, we reach the desired point B. If we desire to avoid this troublesome sudden jump in the current, we may proceed in the following way. The alternator is so excited as to give about the normal voltage of 500 volts at a frequency of 10,200 cycles per second. That is, the set is being worked at the point B'''. Then, by keeping the voltage constant, but gradually increasing the frequency to 10,500, we pass to the point B thru the points B'' and B'.

d. The interesting phenomena represented by the curves of Figure 5 appear only if that part of the inductance which is due to the primary current (that is, the term $-L_1' \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2$), is allowed to become a rather large percentage of the whole inductance. If the assumptions in **c** are changed merely in such a way that for $I_{1(max)} = 100$ amperes, $\frac{N_1 I_{1(max)}}{S}$ is only $\frac{1}{3}$, the characteristic curves are those of Figure 6. They have no falling parts at all, and therefore the conditions are stable everywhere.

e. How far the characteristics are affected by the value of the resistance R_1, is easily deduced from equation (9), and is demonstrated by Figure 7. The curves of this figure have been calculated under the same assumptions as those of Figure 6, but the resistance R_1, instead of being 5 ohms, as in Figure 6, was assumed to be 1 ohm.

V. THE SECONDARY CURRENT OF A FREQUENCY DOUBLER

In considering the secondary circuit of the arrangement shown in Figure 3, we shall restrict ourselves to the case where the secondary coils are connected in opposition and therefore, according to *IIIb*, a doubling of the frequency is effected.

As to the primary circuit the assumption made in IV may still hold. The assumptions to be made for the secondary circuit correspond to those made for the primary. The secondary circuit may be a condenser circuit (Figure 3) and so adjusted that it is at least approximately in resonance to twice the frequency of the alternator. These assumptions being made, we

may neglect all higher harmonics in the secondary circuit and may express the secondary current i_2 by the equation

$$i_2 = I_2 \sin(2\omega t + a).$$

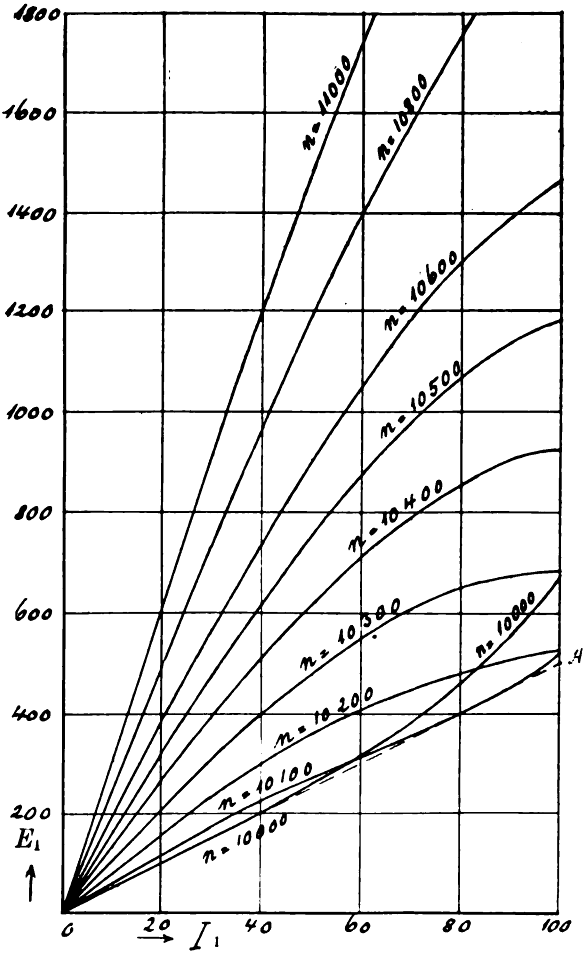

FIGURE 6

The differential equation for the secondary circuit is

$$R_2 \frac{d\,i_2}{d\,t} + L_2 \frac{d^2\,i_2}{d\,t^2} + N_2 \frac{d^2}{d\,t^2}(\Phi_a - \Phi_b) + \frac{i_2}{C_2} = 0 \cdot \quad \cdot \quad \cdot \quad \cdot \quad \cdot \ (13)^*$$

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*L_2 being the inductance of the secondary circuit outside of the frequency changer.

The terms Φ_a and Φ_b have to be calculated from equation (4), where

$$\left.\begin{aligned} F_a &= N_1 i_1 + N_o I_o + N_2 i_2 \\ F_b &= N_1 i_1 - N_o I_o - N_2 i_2, \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (14)$$

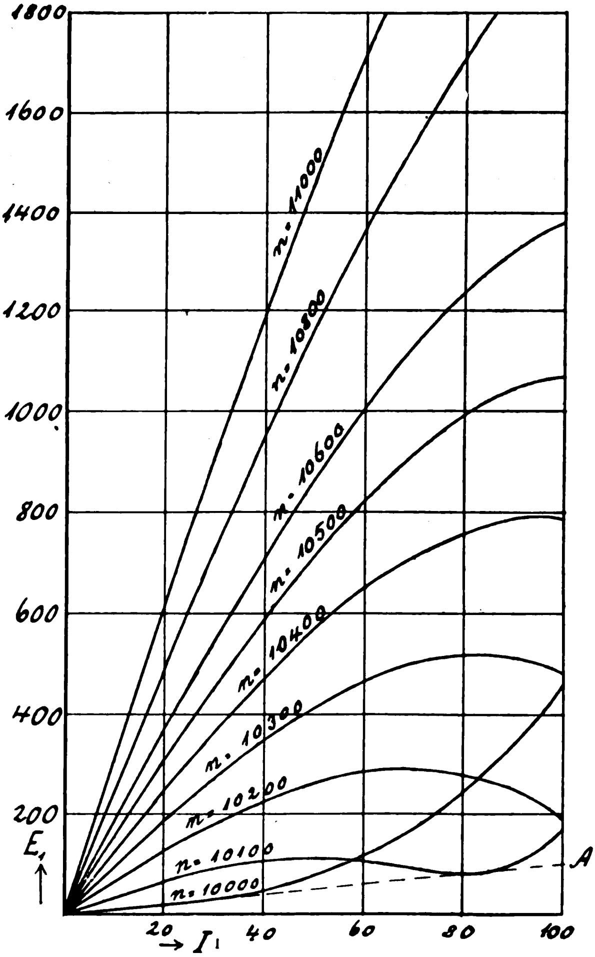

FIGURE 7

since an appreciable secondary current may now be flowing. Substituting these terms in equation (4), we get

$$\begin{aligned}\Phi_a-\Phi_b=&\,2p\,N_oI_o\left\{1-\left(\frac{N_1i_1}{S}\right)^2-\left(\frac{N_2i_2}{S}\right)^2-\frac{1}{3}\left(\frac{N_oI_o}{S}\right)^2\right\}\\&+2p\,N_2i_2\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\left(\frac{N_1i_1}{S}\right)^2-\frac{1}{3}\left(\frac{N_2i_2}{S}\right)^2\right\}\\=&\,2p\,N_oI_o\left\{1-\left(\frac{N_1I_1\sin\omega t}{S}\right)^2-\left(\frac{N_2I_2\sin(2\omega t+a)}{S}\right)^2\right.\\&\left.-\frac{1}{3}\left(\frac{N_oI_o}{S}\right)^2\right\}\\&+2p\,N_2I_2\sin(2\omega t+a)\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\left(\frac{N_1I_1\sin\omega t}{S}\right)^2\right.\\&\left.-\frac{1}{3}\left(\frac{N_2I_2\sin(2\omega t+a)}{S}\right)^2\right\}\\=&\,2p\,N_oI_o\left\{1-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2+\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2\cos 2\omega t-\frac{1}{2}\left(\frac{N_2I_2}{S}\right)^2\right.\\&\left.-\frac{1}{3}\left(\frac{N_oI_o}{S}\right)^2\right\}\\&+2p\,N_2I_2\sin(2\omega t+a)\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2\right.\\&\left.-\frac{1}{4}\left(\frac{N_2I_2}{S}\right)^2\right\}\\&+\frac{p}{2}\cdot N_2I_2\cdot\left(\frac{N_1I_1}{S}\right)^2\sin a\end{aligned}$$

This gives

$$\begin{aligned}N_2\frac{d^2}{dt^2}(\Phi_a-\Phi_b)=&-4\,\omega^2p\,N_2\,.\,N_oI_o\cdot\left(\frac{N_1I_1}{S}\right)^2\cos 2\,\omega t\\&-8\,\omega^2p\,N_2^2I_2\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2\right.\\&\left.-\frac{1}{4}\left(\frac{N_2I_2}{S}\right)^2\right\}\sin(2\,\omega t+a)\end{aligned}$$

and by substituting this expression in equation (13), we easily get

$$R_2I_2=2\,\omega\,p\,N_2\,.\,N_oI_o\,.\left(\frac{N_1I_1}{S}\right)^2\cdot\cos a,$$

$$\left[2\,\omega\left(L_2+L_2'\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2-\frac{1}{4}\left(\frac{N_2I_2}{S}\right)^2\right\}\right)-\frac{1}{2\,\omega C_2}\right]I_2$$
$$=-2\,\omega\,p\,N_2\,.\,N_oI_o\,.\left(\frac{N_1I_1}{S}\right)^2\sin a\quad\cdot\quad\cdot\quad\cdot\qquad\textbf{(14a)}$$

or

$$I_2=\frac{E_2}{\sqrt{R_2^2+\left[2\omega\left(L_2+L_2'\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2-\frac{1}{4}\left(\frac{N_2I_2}{S}\right)^2\right\}\right)-\frac{1}{2\omega C_2}\right]^2}}\tag{15}$$

$$\tan\alpha=-\frac{2\omega\left(L_2+L_2'\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2-\frac{1}{4}\left(\frac{N_2I_2}{S}\right)^2\right\}\right)-\frac{1}{2\omega C_2}}{R_2}\tag{15a}$$

where E_2 has the value given by equation (7) and $L_2'=2pN_2^2$.

Taking into consideration that, for a given primary current, the term $\frac{1}{2}\left(\frac{N_1I_1}{S}\right)^2$ in the denominator of equation (15) is constant as well as $\left(\frac{N_oI_o}{S}\right)^2$, and comparing equation (15) with the corresponding equation (9) for the primary circuit, it is easy to see that both of these equations are of quite the same form. Therefore all that has been said in discussing equation (9) under heading (IV), holds good for this equation also.

VI. THE LOADED FREQUENCY DOUBLER

In IV, the primary current had been calculated for an unloaded frequency changer; that is, on the assumption that $i_2=0$. In V the secondary circuit of a frequency doubler has been considered under the condition that the primary current be given. The loaded frequency doubler which is now to be considered is the case in which the secondary circuit is closed, and the only quantity given for the primary circuit is the emf. e_1 of the alternator. In this case the differential equations (8) and (13) hold at the same time, and for both circuits the number of ampere-turns is determined by equation (14).

Substituting the terms of equation (14) in equation (4), we get

$$\Phi_a+\Phi_b=2pN_1i_1\left\{1-\frac{1}{3}\left(\frac{N_1i_1}{S}\right)^2-\left(\frac{N_oI_o}{S}\right)^2-\left(\frac{N_2i_2}{S}\right)^2-2\left(\frac{N_oI_o}{S}\right)\left(\frac{N_2i_2}{S}\right)\right\}.$$

Under the same assumptions which had been made above for the primary and secondary circuits, this expression is easily transformed into

$$\Phi_a + \Phi_b = 2\,p\,N_1\,I_1 \left\{ 1 - \left(\frac{N_0 I_0}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2 - \frac{1}{2}\left(\frac{N_2 I_2}{S}\right)^2 \right\} \sin \omega t$$

$$- 2\,p\,N_1 I_1 \cdot \frac{N_0 I_0}{S} \cdot \frac{N_2 I_2}{S} \cdot \cos(\omega t + \alpha)$$

Introducing this term into the differential equation (8) for the primary circuit, its solution becomes

$$\left.\begin{aligned} I_1 &= \frac{E_1}{\sqrt{(R_1 + R_1')^2 + \left[\omega\left(L_1 + L_1''\right) - \frac{1}{\omega C_1}\right]^2}} \\ \tan \phi &= \frac{\omega\,(L_1 + L_1'') - \frac{1}{\omega C_1}}{R_1 + R_1'} \end{aligned}\right\} \quad \ldots\ldots \quad (16)$$

where

$$R_1' = \omega L_1' \cdot \left(\frac{N_0 I_0}{S}\right) \cdot \left(\frac{N_2 I_2}{S}\right) \cdot \cos \alpha \cdot \quad \ldots\ldots \quad (17)$$

$$L_1 = 2\,p\,N_1^2$$

$$L_1'' = L_1' \left\{ 1 - \left(\frac{N_0 I_0}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2 - \frac{1}{2}\left(\frac{N_2 I_2}{S}\right)^2 + \left(\frac{N_0 I_0}{S}\right)\left(\frac{N_2 I_2}{S}\right) \sin \alpha \right\} \cdot \quad \ldots\ldots \quad (18)$$

The additional resistance term R_1' is caused by the energy absorption in the secondary circuit. It is easy to show that $R_1' I_1^2$ is identical with $R_2 I_2^2$. Owing to this term the apparent total resistance of the primary circuit $R_1 + R_1'$ is no longer constant, but depends on the value of I_2 and α, and therefore on the load of the frequency changer. The terms in L_1'', namely:

$$-\frac{1}{2}\left(\frac{N_2 I_2}{S}\right)^2 + \left(\frac{N_0 I_0}{S}\right)\left(\frac{N_2 I_2}{S}\right) \sin \alpha,$$

are due to the reaction of the secondary circuit on the magnetic field of the frequency changer.

As for the secondary circuit, the solution of the differential equation (13) is the same as that given in V, namely equation (15). But since the value of I_1 in this equation is not given in this case, but has to be calculated from equation (16), and since on the other hand, equation (16) contains I_2 which is to be derived from equation (15), the situation is far more complicated than in §§ IV or V. It is indeed scarcely possible from the two equations (15) and (16) to get in a general way a clear idea of what really happens. It may be well therefore to restrict ourselves to the discussion of two special cases.

a. The first case may be defined by the condition that the secondary circuit is always adjusted in resonance to twice the frequency of the alternator.

According to this assumption, and to equations (15) and (15a), we get

$$I_2 = \frac{E_2}{R_2}\,; \quad \alpha = 0,$$

and from equations (17) and (18),

$$R_1' = \omega L_1' \cdot \left(\frac{N_o I_o}{S}\right)\left(\frac{N_2 I_2}{S}\right). \quad . \quad . \quad . \quad . \qquad (17a)$$

$$L_1'' = L_1'\left\{1 - \left(\frac{N_o I_o}{S}\right)^2 - \frac{1}{4}\left(\frac{N_1 I_1}{S}\right)^2 - \frac{1}{2}\left(\frac{N_2 I_2}{S}\right)^2\right\}. \quad . \quad . \quad (18a)$$

According to equation (7) E_2 being proportional to I_1^2, the term $\left(\frac{N_2 I_2}{S}\right)$ becomes proportional to I_1^2 and $\left(\frac{N_2 I_2}{S}\right)^2$ proportional to I_1^4.

I have calculated the characteristics of the primary circuit with the same values as those on which the calculation of the curves in Figure 7 was based; and assuming further that

$$\frac{N_2 I_{2(max)}}{S} = \frac{1}{3}, \quad R_1'{}_{(max)} = 4\, R_1,$$

$I_{2(max)}$ being the secondary current produced by a primary current $I_{1(max)} = 100$ amperes at a frequency of 11,000 cycles per second and $R_1'{}_{(max)}$ being the value of R_1' under the same conditions.

The curves are reproduced in Figure 8. They show, that by the reaction of the secondary current, conditions are created which resemble very much those discussed in § IV. I therefore need not go into any details.

Supposing the normal voltage E_1 of the alternator to be about 500 volts, the most favorable conditions are represented by the part *AB* of the curve, corresponding to the frequency 10,300 cycles per second; the equilibrium is stable, and for the normal voltage, the primary, and therefore the secondary current, and the output of the frequency changer are remarkably large.

The efficiency η of the frequency doubler is equal to the output divided by the total input. Therefore

$$\eta = \frac{R_1'}{R_1 + R_1'},$$

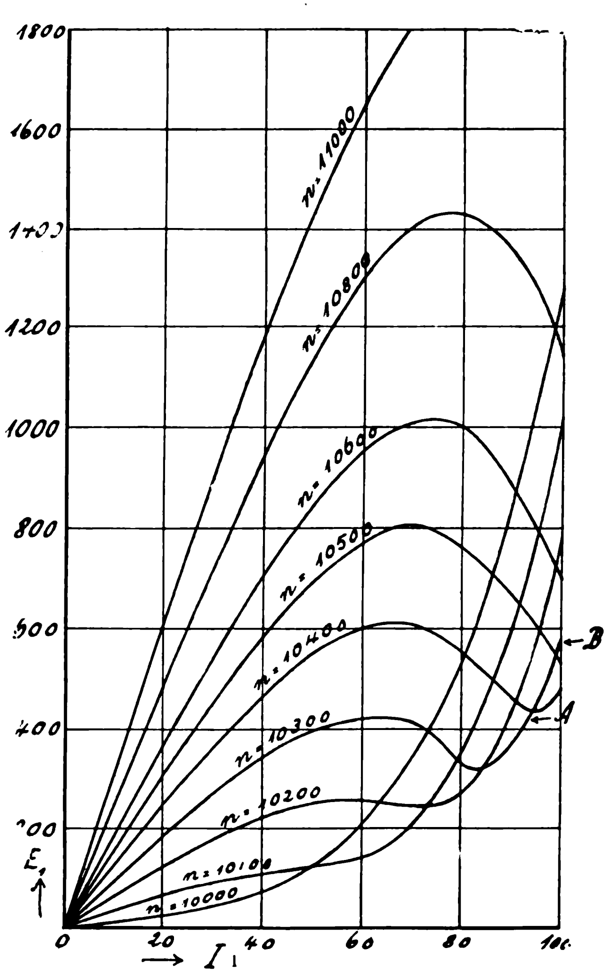

FIGURE 8

and according to equation (17a) it depends largely on the value of I_2 and therefore of I_1. Supposing that for a given frequency and for the current I_2 corresponding to a primary current $I_{1(max)} = 100$ amperes, $R_1' = k \cdot R_1$, we get

$$\eta = \frac{k \cdot \left(\frac{I_1}{100}\right)^2}{1 + k\left(\frac{I_1}{100}\right)^2}.$$

The values of η which correspond to different primary currents and to values of k from 1 to 5, are represented in Figure 9.

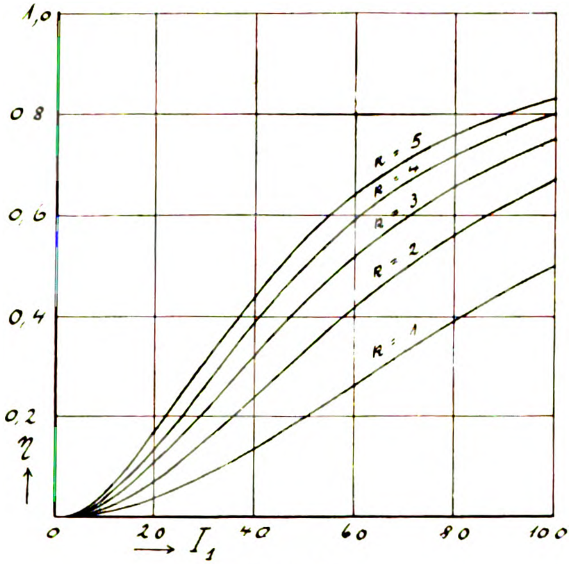

FIGURE 9

b. The second case is based on the assumption that the secondary circuit is adjusted for resonance to one frequency, and then is not varied again. The question to be considered is to what extent the characteristics are changed when the frequency of the alternator is varied.

As soon as the secondary circuit is out of resonance to twice the frequency of the primary current, in equations (17) and (18) a is no longer equal to zero, and in the expression for L_1'' (equation (18)), the term $+\left(\frac{N_o I_o}{S}\right) \cdot \left(\frac{N_2 I_2}{S}\right) \sin a$ no longer disappears, but may have a considerable influence on the value of L_1''. In order to diminish the mathematical difficulties of the problem, the inductance L_2 of the secondary circuit may be assumed to be so large that in the expression for I_2, (equation (5)), the terms containing the primary and secondary current compared with L_2 are too small to change the whole inductance of the secondary circuit materially. Then the conditions are substantially those of a condenser circuit with a constant inductance and therefore, as is well known, the equations (15) and (15a) can be replaced by

$$I_2 = \frac{E_2}{R_2\sqrt{1+\frac{x}{(d/2\pi)^2}}}; \quad \tan\alpha = -\frac{x}{d/2\pi},$$

where d is the logarithmic decrement of the secondary circuit, and

$$x = \frac{n-n_r}{n},$$

n_r being the resonance frequency.

The decrement d of the secondary circuit may be 0.05, and the frequency n_r (to which the secondary circuit is in resonance) may be $=2\times10{,}300$ cycles per second. The assumptions made above regarding the primary and secondary circuit may still hold. Then we get the curves of Figure 10 as the char-

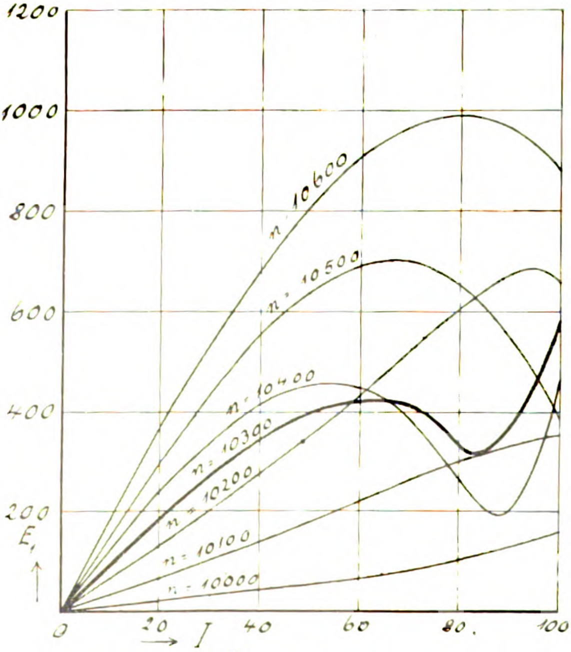

FIGURE 10

acteristics of the primary circuit, according to equations(16), (17) and (18).

These curves show one remarkable feature. In the expression for L_1'' in equation (18) the term $\left(\frac{N_0 I_0}{S}\right) \cdot \left(\frac{N_2 I_2}{S}\right) \sin \alpha$ is positive for frequencies below the resonance frequency and negative for frequencies above it, and under the assumptions made above, this term has a marked influence on the values of L_1''. Owing to this fact the curves for frequencies above the resonance frequency and those below it have quite a different character. Therefore an increase of the frequency of the alternator by 1 per cent. affects the operating conditions in a quite different manner from a decrease of the frequency by the same amount.

VII. INFLUENCE OF THE HYSTERESIS EFFECT

All previous considerations have been based upon the magnetisation curve Figure 2.

a. There is an obvious objection to the use of this curve; namely, that the real magnetisation curve ought to show the hysteresis effect. According to the experiments of Fassbender and Hupka * on the behaviour of iron, we might expect a magnetisation curve like Figure 11 rather than one like Figure 2.

As far as the primary circuit of the unloaded frequency doubler is concerned, this magnetisation curve can easily be represented by an analytical expression which still allows the integration of the differential equation. The magnetic fluxes Φ_a and Φ_b corresponding to Figure 2 are of the form:

$$\Phi_a = U + V \sin \omega t$$
$$\Phi_b = U' + V' \sin \omega t,$$

which may be derived from equations (4) and (5) by neglecting all harmonic terms. In order to get the magnetisation curve for Φ_a of Figure 11 we have only to add the term

$$- r V \cos^3 \omega t \, (1 - \sin \omega t),$$

where in the case of Figure 11, $r = 0{,}45$. We therefore get

$$\Phi_a = U + V \left[\sin \omega t - r \cos^3 \omega t \, (1 - \sin \omega t)\right].$$

The corresponding equation for Φ_b is

$$\Phi_b = U' + V' \left[\sin \omega t - r \cos^3 \omega t (1 + \sin \omega t)\right].$$

---

* H. Fassbender and E. Hupka, "Jahrbuch der drahtlosen Telegraphie," 6, page 133, 1912.

In consequence of these, we have

$$N_1(\Phi_a+\Phi_b)=A+B\left[\sin\omega t-\frac{3}{4}r\cos\omega t\right], \qquad (19)$$

again neglecting all harmonic terms. Comparing this expression with that calculated in III*a*, we find

$$B=L_1'\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{4}\left(\frac{N_1I_1}{S}\right)^2\right\}I_1.$$

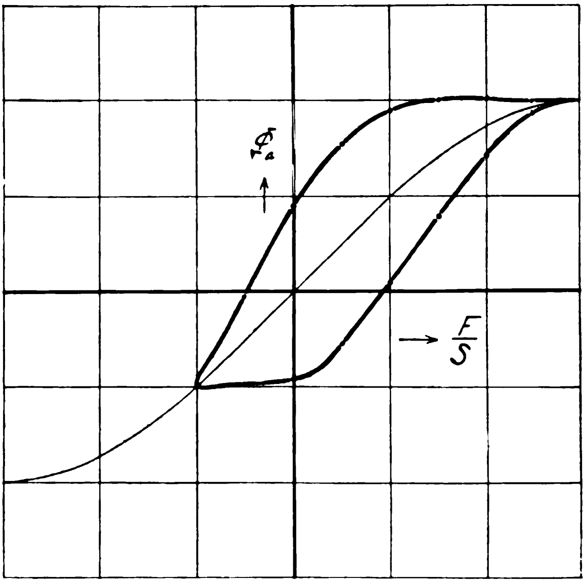

FIGURE 11

Substituting this in equation (8) for the primary circuit of the unloaded frequency doubler, we obtain the result that the conditions in the primary circuit are changed by the additional term $-\frac{3}{4}Br\cos\omega t$ only in so far as there is added to the resistance R_1 a term R_1'' where

$$R_1''=+\frac{3}{4}\omega rL_1'\left\{1-\left(\frac{N_oI_o}{S}\right)^2-\frac{1}{4}\left(\frac{N_1I_1}{S}\right)^2\right\}$$

This term, expressing the iron losses, depends on the primary current. But even assuming $\frac{N_1I_{1(max)}}{S}$ to be $=\frac{2}{3}$ (compare **IVc**) for $I_{1(max)}=100$ amperes, the variation of the value of this term is not more than about 11 per cent., when I_1 is changed from 0 to 100 amperes.

The conditions for the loaded frequency doubler become somewhat more complicated, but it is easy to show that the results obtained in VI are not changed materially.

b. A by far more serious objection may be raised against the harmless looking equation (3)

$$\left.\begin{matrix} F_a \\ F_b \end{matrix}\right\} \leqq S$$

This equation, which is absolutely necessary under the assumption of equation (2), does not allow an increase of the primary and secondary currents such as would make the number of ampere turns greater than that corresponding to the point A in Figure 1. Now, a real magnetisation curve—not taking into account the hysteresis effect—looks rather like the curve Figure 12. There are good reasons to believe that in many respects the results obtained would be very interesting if we were able actually to increase the ampere turns so as to reach the part AB of the magnetisation curve Figure 12. We are therefore seriously handicapped by equation (3). I shall try to surmount this limitation, and I hope to be able, in a later paper, to report on the results of this later investigation.

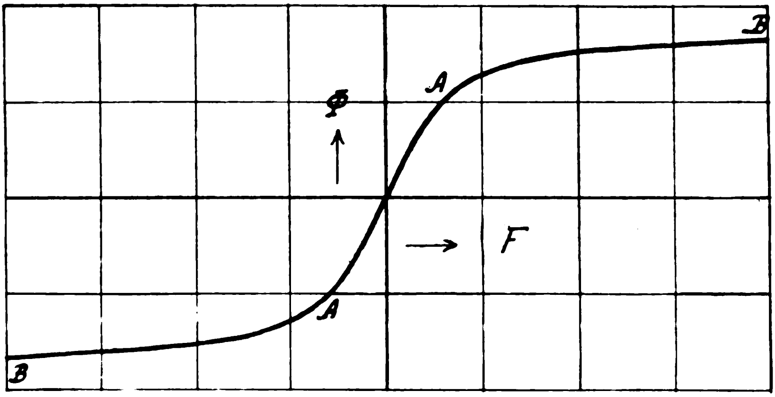

FIGURE 12

SUMMARY: Using iron-core one-to-one transformers, with (or without) supplementary d.c. excitation, the well-known magnetic frequency doublers and triplers can be arranged. The theory of these is developed, assuming a cubic relation between flux and field intensity.

The emf. in the secondary, the effect of supplementary d.c. excita-

tion, the primary current and phase, and the secondary current for the unloaded doubler are derived. The results are extended to the loaded doubler, the efficiency and operating characteristics of which are calculated.

The modification of the derived formulas due to iron hysteresis is obtained.

The theory is clearly illustrated by a number of calculated curves; of which the primary current-voltage curves show an interesting combination of stable (rising) and unstable (falling) portions. The resulting desirable methods of practical operation are fully considered.

The theory is to be further developed to cover phenomena dependent on iron saturation above the knee-point of the magnetization curve.

SOME CHARACTERISTICS OF THE FREQUENCY DOUBLER AS APPLIED TO RADIO TRANSMISSION*

By

T. MINOHARA

(NAVAL ARSENAL, TOKIO, JAPAN)

1. GENERAL THEORY AND VALUE OF SINUSOIDAL EMF. INDUCED IN THE SECONDARY

In the well known Joly-Vallauri system, as shown in Figure 1, we have a pair of transformers T_a T_b excited with direct current A_D, where e_1, e_2 denote the voltages in primary and secondary; e_a, e_b denote the voltages of each transformer T_a, T_b, a_1,a_2 denote the primary currents and secondary currents respectively, and for simplicity, the number of windings are assumed to be all the same and unity.

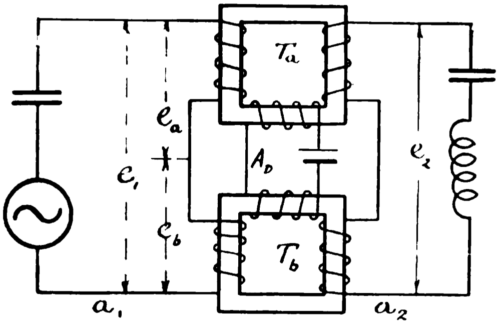

FIGURE 1

The resonance effect being generally utilized in radio engineering to get the maximum output, the wave forms of oscillating current a_1,a_2 can be assumed to be nearly sine curves, or $a_1 = A_1 \sin \omega t$, $a_2 = A_2 \sin 2\omega t$. This assumption makes it easy to solve all problems dealing with the frequency transformer.

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* Received by the Editor, September 28, 1919.

The Lissajous figure, shown in Figure 2, is taken with a Braun tube, the light spot being deflected horizontally by the primary current a_1 of 50,000 cycles and vertically by the secondary current a_2 of 100,000 cycles in a circuit of the type used in radio practice. This figure clearly shows that the wave forms of a_1 and a_2 both are nearly sinusoidal.

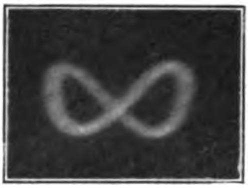


FIGURE 2

The general principle of the frequency doubler is generally assumed to be as follows:

When certain relations exist between A_D and a_1 in the magnetizing current A_D+a_1, the emf. e_a induced in the secondary of transformer T_a would take the form

$$e_a = P_1 \, sin \, \omega t + P_2 \, sin \, 2 \omega t$$

Similarly the emf. e_b induced in the secondary of the other transformer T_b, excited by the current $A_D - a_1$ would be

$$e_b = P_1 \, sin \, (\omega t - \pi) + P_2 \, sin \, (2 \omega t - 2 \pi).$$

If e_a and e_b are superposed, we get

$$e_2 = e_a + e_b = 2 \, P_2 \, sin \, \omega t$$

which is a pure sinusoidal wave of double frequency, the fundamental wave being entirely cancelled out.

The necessary relation of A_D and a_1 required can be easily determined experimentally, but we can also find it graphically by considering the magnetization curve of the iron cores.

Taking the core the magnetization curve of which is shown in Figure 3, if we try to plot the curves of flux and induced emf., assuming arbitrary ampere-turns by trial so as to obtain a curve of secondary induced emf. e_2 of pure sine wave form, we shall find it desirable to take the primary current a_1 so that $A\,B = 1.6$ to $1.8 \; C\,D$. For example, in Figure 3, taking a_1 so that $A\,B = 1.8\,C\,D$, we get the curve ϕ_a and $\frac{d\,\phi_a}{d\,t}$ or e_a as in Figure 4.

Adding e_a and e_b, which have the same wave form with 180° phase, difference we obtain the nearly sinusoidal wave of double frequency, as shown in Figure 5.

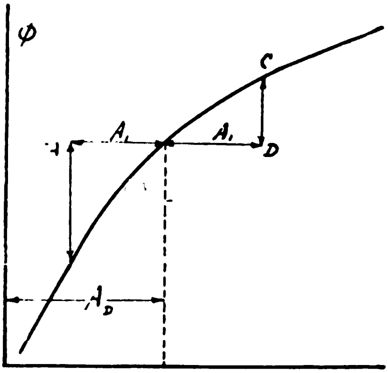


FIGURE 3

From my calculations given below, the relation $AB = 1.667\,CD$ was deduced for the same condition.

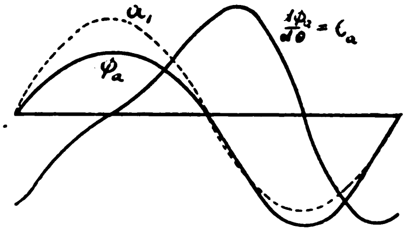


FIGURE 4

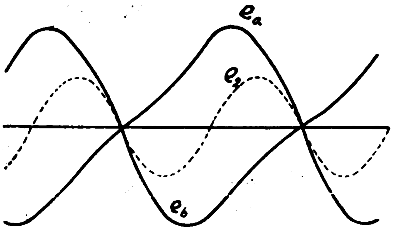


FIGURE 5

In the ideal case, where no useless higher harmonics are considered, the flux in each transformer core would be

$$\phi_a = -\int e_a\,dt = -\int (P_1 \sin \omega t + P_2 \sin 2\,\omega t)\,dt$$

$$= \frac{P_1}{\omega} \cos \omega t + \frac{P_2}{2\,\omega} \cos 2\,\omega t$$

and

$$\phi_{\omega t=0} = \frac{P_1}{\omega} + \frac{P_2}{2\,\omega}.$$

$$\phi_{\omega t=\pi} = -\frac{P_1}{\omega} + \frac{P_2}{2\,\omega}.$$

Taking

$$\frac{P_1}{P_2} = k,$$

we get the relation

$$\frac{\phi_{\omega t=0}}{-\phi_{\omega t=\pi}} = \frac{2\,k+1}{2\,k-1} = X.$$

We now plot the curves e_a', e_a'', e_a''' in Figure 6, assuming

$$e_a' = P_1 \sin \omega t + P_2' \sin 2\,\omega t$$

$$e_a'' = P_1 \sin \omega t + P_2'' \sin 2\,\omega t$$

$$e_a''' = P_1 \sin \omega t + P_2''' \sin 2\,\omega t$$

where $$P_2''' < P_2'' < P_2'$$

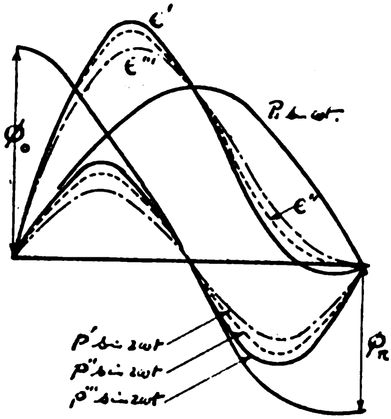


FIGURE 6

In order that the amplitude of the second harmonic should be as great as possible, the curve e_a' is the best of the three given. But it is impossible that the curve e_a pass thru the zero line during the interval from $\omega t = C$ to π, as e_a' does, since $\frac{d\phi_a}{dt}$ can never be zero during that time, as seen from Figure 6. Therefore, in the curve e_a, which would give the maximum realisable second harmonic, $P_2 \sin 2\,\omega t$, the value $\frac{d\phi}{dt}$ should be equal to zero at $\omega t = \pi$.

or

$$\left(\frac{de}{dt}\right)_{\omega t = \pi} = 0,$$

$$[\omega P_1 \cos \omega t + 2\,\omega P_2 \cos 2\,\omega t]_{\omega t = \pi} = 0.$$

Then we have $\quad P_1 = 2P_2, \quad$ or $\;k = 2.$

Substituting this value in the condition X above, we get

$$X = \frac{\phi_{\omega t = 0}}{-\phi_{\omega t = \pi}} = \frac{2k+1}{2k-1} = 1.66.$$

This means that $AB = 1.667\,CD$.

Let us call the case of the secondary induced emf. of pure sinusoidal wave form, a "sine wave system."

2. IMPULSIVE EMF. INDUCED IN THE SECONDARY.

In the sine wave system a_1 and e_2 can not be great enough to get the sufficient secondary power, owing to the limitation $AB = (X)\,(CD)$, mentioned above.

In practice, it is usually required to get as much energy as possible, the induced emf. e_2 being used for impulse excitation of the secondary oscillation, and the second harmonic only being effectively resonated by proper choice of the secondary circuit frequency. Therefore it is not essential that the secondary induced emf. be a pure sinusoidal wave of double frequency, but it is usually necessary to obtain the maximum impulsive secondary emf. of double frequency. Thus the flux density change due to a_1 should be large in order to get a large secondary induced emf. for a given size of core. That is, a larger a_1 is required as compared with A_D than for the case of the sine wave system.

Taking $A_1 > A_D$ in the example shown in Figure 7, the effective value of a_1 to be 10 amperes (or $\sqrt{2} \times 10$ as the maximum amplitude A_1), and $A_D = 5$, then the corresponding flux curve can be plotted as shown in the curve ϕ_a (assuming that its saturation curve be as shown in Figure 8) and we can get the curve e_a from the inclination of ϕ_a or $\frac{d\phi_a}{dt}$. Similarly the curve e_b is obtained, which has the same wave form, but with a phase difference of 180° from e_a. The secondary voltage of double

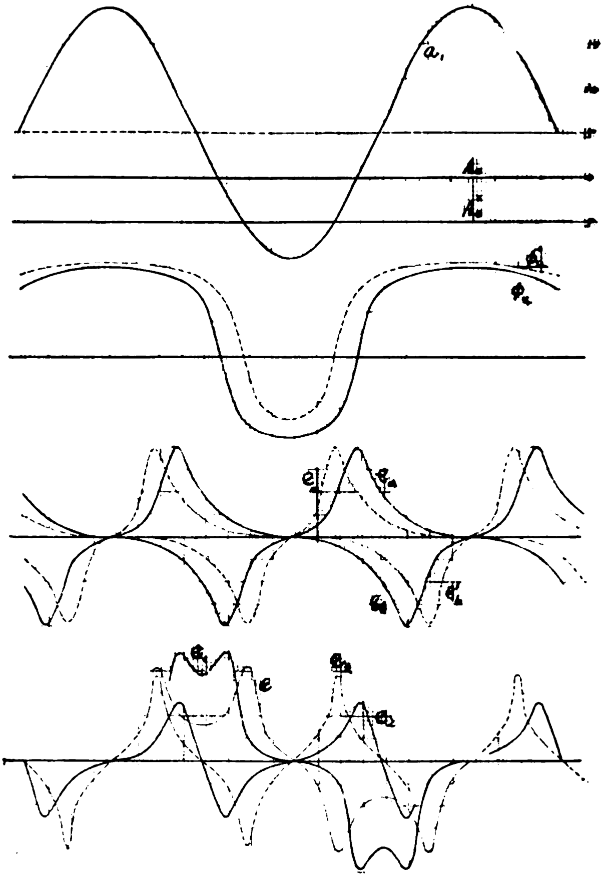


FIGURE 7

frequency will be the sum of e_a and e_b, and the primary voltage e_1 is the difference of e_a and e_b, as shown in Figure 7. But if the excitation $A_D' = 10$ (or effective value of a_1), the flux curve will be ϕ_a', then e_a', e_b' are as plotted, and $e_a' + e_b' = e_2'$ and $e_a' - e_b' = e_1'$ will be as shown in the figure.

As seen in the figure, the wave forms e_a and e_b' both are not sinusoidal but sharp-peaked or impulsive waves, containing many higher harmonics. The maximum amplitude of e_2 does not come at equal intervals while that of e_2' does come with a regular period of double frequency. The amplitude of e_2' is much greater than that of e_2, therefore we can say that A_D should be equal to the effective value of a_1 in order to get the maximum impulsive emf. of double frequency.

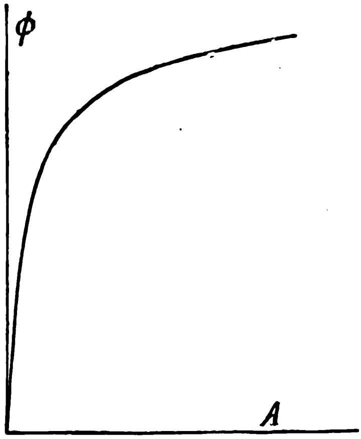


Figure 8

The fact is explained as follows:

The amplitude of e_2 will be the maximum at the same phase as the maximum amplitude of e_a and e_b or $\frac{d\phi_a}{dt}$. But $\frac{d\phi_a}{dt}$ should be the maximum when the excitation current $A_D \pm A_1 \sin \omega t$ in the cores becomes zero, if we neglect the effect of the hysterisis loop. Hence $A_D \pm A_1 \sin \omega t$ should be zero at the phase of $\frac{\pi}{4}$ or $\frac{3}{4}\pi$ in order to produce the maximum amplitude of e_a at intervals of $\frac{\pi}{2}$ (or π of double frequency) as seen in Figure 7.

We have
$$A_D - A_1 \sin \frac{\pi}{4} = 0$$
or
$$A_D = \frac{A_1}{\sqrt{2}}.$$

This means that d.c. ampere-turns are equal to the effective value of primary ampere-turns.

If A_D be increased to the point of magnetic saturation, always maintaining the relation $A_1 = \sqrt{2}A_D$, the curve e_a becomes gradually a peaked wave, and the fundamental wave tends to disappear, the position of the maximum amplitude being unchanged.

As an example ϕ', ϕ'', ϕ''' in Figure 9 show the flux curve taking $A_D = \frac{A_1}{\sqrt{2}} = 5$, $A_D = \frac{A_1}{\sqrt{2}} = 10$, $A_D = \frac{A_1}{\sqrt{2}} = 15$ respectively

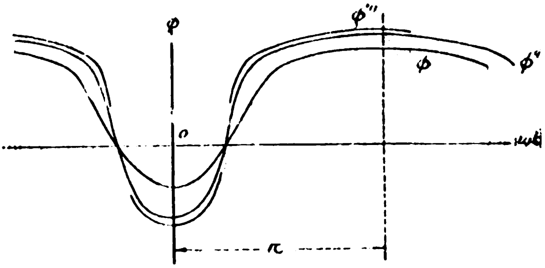


Figure 9

in the saturation curve shown in Figure 8. Furthermore, the corresponding induced emf. e_a', e_a'', e_a''' become as shown in Figure 10. Thus if A_D excite the core near to magnetic saturation, the curve e_a become a somewhat peaked wave which consists mainly of the second harmonic, and there is almost no fundamental wave

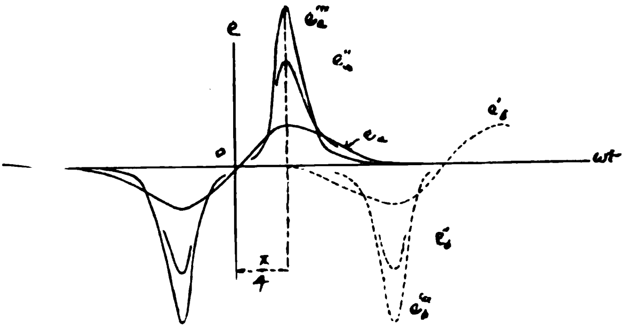


Figure 10

component which will have to be cancelled out when e_a is superposed on e_b to get e_2 (see e_a'', e_b'' or e_a''', e_b'''). If A_D excite the core above the knee of the saturation curve, the curve e_a becomes too peaked and narrow, (see the curve e_a''') and the second higher harmonic also tends to disappear.

Therefore it is preferable that the difference of flux change due to $+A_1$ and $-A_1$ shall be as great as possible, taking A_D so that the excitation is at the knee of the saturation curve, (see the curve e_a'' where $A_D=10$; that is A_D is the excitation corresponding to the knee of the saturation curve).

Thus the two conditions $A_D=\frac{A_1}{\sqrt{2}}$ and A_D equal to the excitation corresponding to the knee of the saturation curve are necessary to get the maximum impulsive wave of double frequency.

I will call this system the "impulsive wave system." It is important in practice at the lower radio-frequencies and is discussed in the following sections.

Let us consider a special imaginary case where the magnetic flux of the core increases linearly with excitation up to the knee of the saturation curve (for example, with an excitation of 4 amperes), and past this point, the flux maintains a constant value, as shown in Figure 11.

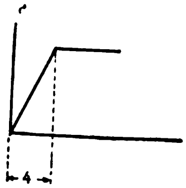


FIGURE 11

Taking $A_D=4$, and the effective value of the primary current (the wave form of which is a pure sine) as 4 amperes, if we plot the flux curve corresponding to the excitation a_1+A_D and a_1-A_D, we get the curves of ϕ_a and ϕ_b as shown in Figure 12. If, then we plot the curves e_a and e_b as calculated from the slope of ϕ_a and ϕ_b, we can see from these curves that the effective value of e_2 in this case is equal to the sum of effective values of e_a and e_b, that the effective value of e_2 is equal to the effective value of e_1, that the wave form of e_1 is sinusoidal, and that the form factor of e_a is equal to that of a sine wave.

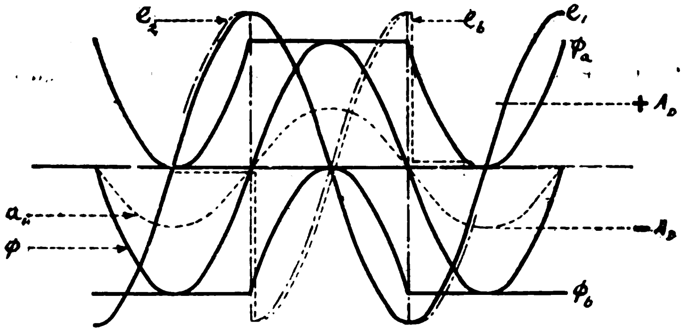


FIGURE 12

3. DISTORTION OF VOLTAGE WAVE DUE TO THE LOAD CURRENT

We will explain the effect of loading the frequency changer, using as an example the core the characteristic of which is shown in Figure 8.

The secondary voltage at no load can be plotted as shown in Figure 13. If the primary current is a_{1_0} and exciting current A_D, then the flux curve will be ϕ_0 in Figure 14, and the terminal voltage of each transformer will be e_{a_0}. Therefore we can easily obtain the secondary voltage by adding the ordinates of e_b which have the same wave form but with 180° phase difference.

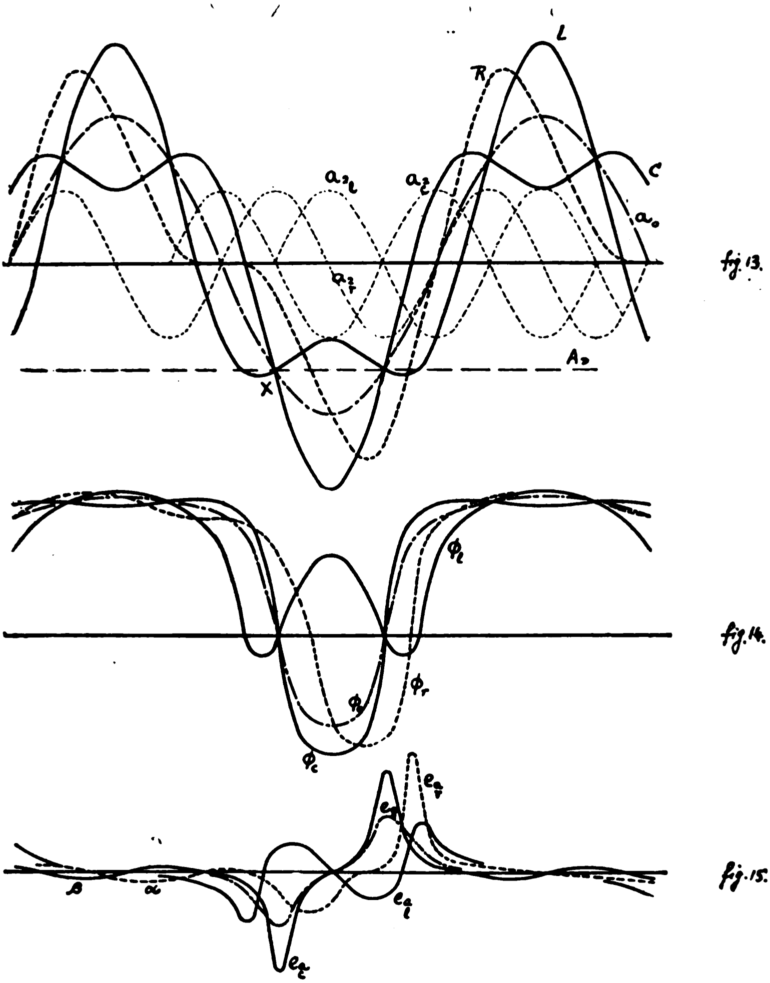
(a). INDUCTIVE LOAD

On loading the changer inductively, the secondary current lags behind the emf. Taking the curve of the secondary inductive current a_{2l} in Figure 13, which lags 90 degree behind the emf. of no load, the maximum amplitude of which should be at the phase of X in the figure, we can plot the corresponding flux curve of a transformer the exciting current of which is $a_{2l}+a_{1_0}+A_D$, or $L+A_D$, as shown in Figure 14. Thus we get quite an irregular wave form e_{al} by calculating $\frac{d\phi_l}{dt}$ from the curve ϕ_l as shown Figure 15. Therefore, if we try to plot the wave form of e_2 by adding e_{a_0} and e_{b_0}, the wave forms of which are the same but with 180° phase difference, we can easily see that an irregular wave form is obtained, and its amplitude is comparatively low, hence the output can not be great in that case.

(b). CAPACITY LOAD

Assuming secondary capacity load current a_{2_c}, with phase 90° ahead of the secondary emf. at no load, the sum of a_1 and a_{2_0}

will be C, as shown in Figure 13. Then, if we plot the flux curve of a transformer corresponding to the exciting current a_1+C or $a_1+a_{2_b}+A_D$, we get the curve ϕ_c as shown in Figure 14.



FIGURES 13 to 15

Similarly we can get e_{a_c} as shownin Figure 15 with maximum amplitude far greater than e_{a_c}. So if we plot e_2 by adding e_{a_c} and e_{b_c}, we find that the amplitude of e_2 is great and the wave has quite a regular form of double frequency where the part β of the curve is obtained when e_{a_c} and e_{b_c} are superposed to get e_2.

These capacity or inductive loads dissipate practically no power since the phase difference of a_2 and e_2 is nearly equal to 90°. Practical operation should always lie between these two, the resonance effect being employed.

(c) ACTUAL CASE OF LOADING

If the maximum amplitude of secondary induced emf. were always at the phase of the point X in Figure 13, the secondary current a_2 should also be in the same phase when the secondary circuit is just at resonance condition. But in the actual case, the phase of the induced emf. e_2 will shift by a certain amount as soon as the secondary current a_2 flows since the wave form of the flux is distorted by the secondary current. For example, assume a_{2_r} the secondary current as shown in Figure 13. Then the curve $a_1+a_{2_r}$ is represented by R, and ϕ_r will be the flux curve for this case. The voltage curve of the transformer will be e_{a_r}, which has an unsymmetrical form as shown, and in this example the secondary current a_{2_r} is advanced about 60° before the induced emf. e_2.

(d) MAXIMUM OUTPUT

As just mentioned, the induced secondary emf. e_2 is high when the output is supplied to a capacity load. On the other hand, it is evident that the output at the resonance condition is a maximum if the secondary induced emf. is kept constant. Therefore we can say that the maximum output is obtainable when the secondary current leads the emf. by a certain small amount. This amount depends on the relative values of a_1, a_2, A_D, and on the saturation curve of the core. In practice, a variometer in the secondary circuit is adjusted so as to obtain maximum current or maximum output, since the effective resistance of the secondary circuit is constant. Therefore it should be remembered that the current and induced emf. in the secondary, in practical cases are not exactly in phase but that the current somewhat leads the emf. e_2.

We have next to find how many secondary turns are required to get the maximum secondary output. As we can see from Figure 14, the curve of ϕ_r deviates from ϕ_o because of the secondary current, and the greater the secondary current a_{2_r} the more is the flux curve distorted. At the same time the part a of the curve shown in Figure 15 tends to increase in amplitude which is very unfavorable as regards obtaining large values of e_2 since this part of e_{a_r} cancels the other part of e_{b_r} in the secondary and accordingly the amplitude of e_{2_r} can not be sufficiently large.

From these considerations, we can conclude that:

The secondary ampere-turns should lie within such limits that the secondary ampere-turns at full load will not produce any considerable amplitude corresponding to a in Figure 15, because of distortion of the flux on load based on the saturation characteristic curve of the core used.

4. OSCILLOGRAMS OBTAINED IN EXPERIMENTS

As the oscillograph will not correctly trace waves at high frequency, I took many oscillograms during the experiments at 50 cycles, using the core the characteristic of which is shown in Figure 8 and the connection diagram as shown in Figure 1.

CASE 1

At no load, the wave form of e_a, e_b, e_1, and e_2 gave the oscillograms shown in Figure 16 (a_1, e_1). From Figure 17 (e_a, e_b), and Figure 18 (e_1, e_2), we can see:

(1) The primary voltage e_1 has a V-topped wave form.

(2) The secondary voltage of double frequency has a uniform period if the exciting d.c. ampere-turns are equal to the effective value of primary ampere-turns.

(3) The phase difference between e_1 and a_1 is about 90°.

CASE 2

On resistance load, the wave forms of e_1, e_a, e_b, a_1, e_2, and a_2 were taken in the oscillograms as shown in Figure 19 (a_1, e_1), Figure 20 (e_a, e_b), and Figure 21 (a_2,e_2) whence we can see that:

(1) The primary voltage of the transformer is distorted a little and the upper and lower parts have become asymmetrical.

(2) The primary voltage has a flat asymmetrical V-topped wave form.

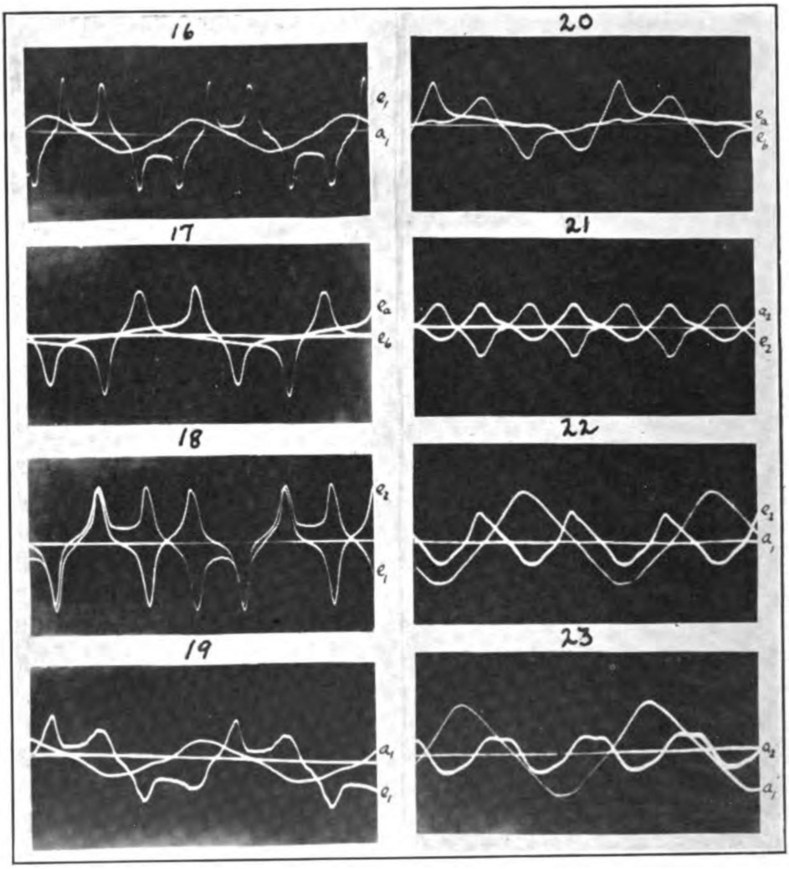
(3) The resultant secondary terminal voltage is somewhat low and asymmetrical.

CASE 3

On capacity, inductance and resistance load, where the variable inductance was adjusted so as to obtain the maximum current, the wave forms a_1, a_2 and the secondary terminal voltage e_2 were taken in oscillograms, as shown in Figure 22 (a_1,e_2), and Figure 23 (a_1,a_2).

This case actually occurs in radio transmitters as already mentioned, and from these oscillograms we can see that the secondary current somewhat leads before secondary terminal voltage e_2.

These results are all clearly explained by the diagrams shown in Figure 13, 14, and 15.



Figures 16 to 23

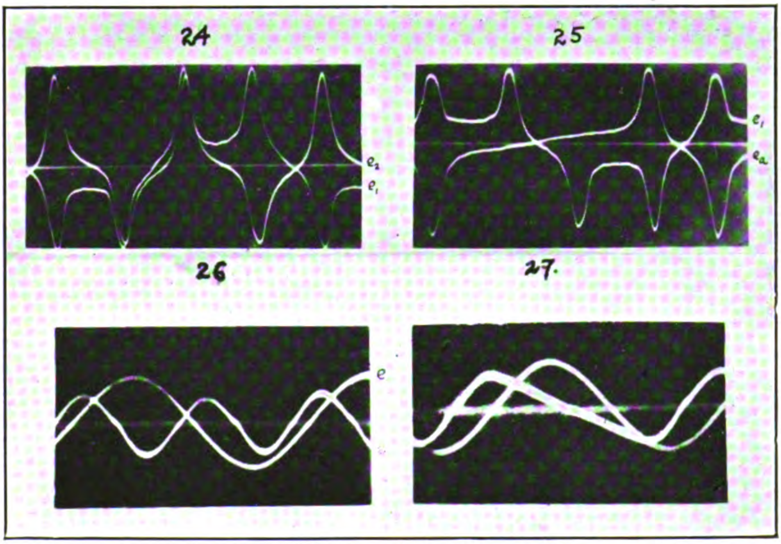
5. Comparison of the Sine Wave System and the Impulsive Wave System

As already mentioned, the secondary emf. of the pure sinusoidal wave obtained by the sine wave system is applied efficiently in producing the secondary output but the flux density change being limited by the condition $AB+X.CD$, the induced emf. or output can not be great as desirable for a given size of core.

Ample space is required for the winding of A_D but the iron loss per unit volume of the core may be small. On the other hand, in the case of the maximum impulsive wave system which has been treated in sections 2 to 4, the flux density change can be made great enough to induce a high secondary emf. because of the necessary conditions that $A_D = A_1/\sqrt{2}$ and $A_D =$ the excitation required to cause the flux just to reach the knee of the saturation curve. Thus the maximum secondary induced emf. wherein many higher harmonics are included, is utilized for impulse excitation in the production of the secondary oscillations wherein the second harmonic predominates because of the resonance effect. Consequently, a satisfactory amount of energy can be drawn from the source. The iron loss per unit volume is great in this case but the total volume of the core may be much less than in the former case.

These two systems correspond closely to the cases of the oscillations of the first type and the second type in Poulsen arcs.

I took oscillograms of e_1, e_a, e_2 for each system adjusting the conditions to the best values; e_1, e_a, e_2 for the impulsive wave system are shown in Figures 24 and 25, and e_1, e_a, e_2 for the sine wave system are shown in Figures 26 and 27. From these curves, we see that the curve e_2 is approximately sinusoidal and the curve e_1 has a flat top in the case of sine wave; while the curve



FIGURES 24 TO 27

e_2 is a peak-topped wave and the curve e_1 has a sharp V-shaped top in the case of impulsive wave.

In practice, it is important to cause the oscillations to have as much energy as possible by using the method of impulsive wave excitation, even tho many higher harmonics are produced, provided the iron loss in the core is of allowable value for the given frequencies. These considerations are similar to those governing the choice of oscillations of the second type in the Poulsen arc. But the sine wave system should be used, if the frequency is too high to enable reasonable losses for a given material of the core.

SUMMARY: A general analysis is given of the theory of the ferromagnetic frequency doubler. The production of secondary circuit energy is secured either by the induction of a sinusoidal secondary emf. or by the production of a sharply peaked, "impulsive" secondary emf. In each case, secondary resonance is employed to emphasize the desired double frequency.

The sinusoidal wave and impulsive wave systems are oscillographically studied and compared for different types of secondary load. The impulsive system is recommended as having higher efficiency wherever the frequency is sufficiently low to avoid excessive iron core losses with that form of wave.

THE STATUS OF THE STATIC FREQUENCY DOUBLER IN RADIO ENGINEERING PRACTICE*

By
FREDERICK C. RYAN
(RADIO ENGINEER)

Further advancement in the art of radio communication calls for the co-ordination of development centralized on the physical phenomena producing the most nearly ideal results so far attained. It is obvious that the method of signal reception based on the interaction of two radio frequency currents properly related to one another, their resultant giving a note of audible frequency, represents our nearest approach to perfection. If both of these currents are of sustained wave form and the receiver is of the vacuum relay type of proper characteristics, the signal note will be purely musical and the pitch subject to variation at will. There is also a marked gain in audio frequency energy by reason of the employment of the received energy in conjunction with that supplied by the locally produced oscillations.

Based on these premises, a consummation of our technical ideals will be fulfilled thru concentration of effort on the development of devices for the most economical generation and radiation of sustained wave radio frequency energy. Ultimate practical perfection requires resourcefulness in reducing energy losses in apparatus and circuits; flexibility of frequency variation; stability of operating frequency; elimination of the undesirable effects of harmonics including those inherent in the generator or induced by reason of the necessarily distributed constants of a radiating circuit; production of sufficient power; and the required speeds of signaling.

Present practice presents as a basis for development the theories applied in the design of radio frequency generators of the direct current arc oscillator, power electron tube, and rotating mechanical types. The arc method involves the passage of an electric current from a copper to a carbon electrode, surrounded by an hydrogenous atmosphere and acted upon by a tranverse magnetic field. The gas serves as a vehicle of de-

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* Received by the Editor November 26, 1917. Bibliography is at end of paper.

ionization of the space between the electrodes by reason of its high coefficient of diffusion and is also beneficial because of its high heat conductivity thus preventing local temperature rises. The magnetic field is an adjunct for ionic removal. By these means rapid growth of the ignition voltage is assured and the apparatus acts as a converter of continuous current energy into sustained radio frequency energy. The application of these hypotheses has afforded us means, at the present time, of producing ample energy for practical wave propagation over normal transmission distances. Rapid manipulation of frequency is obtained by the simple process of varying the antenna system inductance thru appropriate external variable reactance, because the frequency is determined wholly by the circuit constants. Signaling is performed by a small percentage change in radiated frequency, thru the change of inductance by electrical or mechanical methods. It may also be accomplished by shifting of the radio frequency energy from the radiating to a dissipating system. Practical operation of the arc oscillator is only successful when accomplished without disturbance of the factors governing the constancy of the frequency and oscillation amplitude. Therefore the supply circuit must be fixed and signaling done by one of the methods discussed. The inherently unstable characteristics of the arc make it difficult to design absorbing systems and to devise appropriate manipulation thereof to obtain signal control with a minimum effect on the constancy of energy and frequency. Prevalence of harmonics generated and the disturbances caused by the control apparatus detract much from an otherwise desirable generator.

The arc systems losses are somewhat larger than usual, and adding these to those produced by the continuous energy flow at full power causes the operating efficiency to become quite low.

By making use of the properties of free electronic emission from incandescent bodies in vacua, and causing thermionic currents to flow thence to anodes strategically situated we are able to produce radio frequency alternating currents by the employment of a proper control. If a third electrode be brought into the space between such a cathode and anode, the number of electrons returning to the cathode will increase if this body be negatively charged; and, on the other hand, if it be positively charged it will tend to neutralize the negative charges on the electrons in transit. By these artifices the thermionic current is decreased to a minimum or increased to saturation. By introducing into the electrode circuits appropriate inductances and capacities and causing interaction by mutual inductance between them, the regenerative action brought about by the transfer of energy back and forth will cause the device to act as a source of radio frequency alternating current or sustained wave energy. The work already done along these lines promises early fulfillment of the realization of means for accomplishing radio communication with simplicity of signal control, both telegraphic and telephonic, ease of manipulation of energy and high generating efficiency. At this time the high first cost and prohibitive operating expenses exclude their consideration as an economical commercial proposition.

Numerous investigators have engaged in the design of rotating radio frequency alternators. The two types attaining some degree of success are represented by the Alexanderson and the Goldschmidt types of machines. These machines generate and deliver directly to the radiating circuit energy of its fundamental frequency. The Alexanderson alternator[1] has been built and operated in units up to 75 kilowatts and to frequencies of 100,000 cycles. The inductor type of construction is followed in this machine, using a continuous wave armature winding. In order to obtain radio frequencies, it is necessary to have a large number of slots and rotating field projections, together with a high peripheral speed. The output is a function of the air gap clearance because the generated voltage is nearly inversely proportional to the air gap length. Carefully built flexible shafts are necessary because of the high speeds involved, in order that the rotating member may revolve about its exact center of mass. The mechanical design, selection of materials and assembly of a machine of this type involves considerable expense if successful practical operation is to be expected. The windage and friction losses are quite high thereby giving relatively low efficiencies. The question of speed regulation is all important in order to secure stability at the operating frequency. Flexibility of frequency change brings up intricate speed design questions. For medium energy output at not too high a frequency, and where wave length changes are not often required, this apparatus will doubtless find a field of application. The control of such a generator for either telephonic or telegraphic use has been beautifully worked out by Mr. Alexanderson.*

The Goldschmidt type of alternator[2] may be described as a combined generator and frequency multiplying device. By appropriate choice of circuit constants, various tuned circuits are shunted across the stator and rotor windings, serving to intensify and increase the higher harmonics of the generated frequency. That higher harmonic desired for use appears in a resonant radiating circuit. Each of these reflection steps involve losses in machine and circuits, hence it is desirable to reach the requisite frequency with as few steps as possible. Therefore, the machine is designed for a high initial frequency and speed. The mechanical design is exceedingly complex, very small air gaps being necessary. The flux carrying structure is of very thin steel laminations, separated by paper for insulation against eddy currents. As the rotor carries conductors, great care is necessary to secure proper mechanical strengths for successful operation at high speeds. Abundant power is made available by this design, signaling being facilitated thru controlling the direct current exciting field. Wave length change is obviously somewhat difficult to carry out, necessitating cumbersome mechanism.

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*See PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, Volume 4, number 2, page 101, April, 1916.

The construction and operation of alternators giving frequencies up to 10,000 cycles in large power units and as great as 20,000 cycles in small sizes is a practical commercial proposition. By using the inductor type of machine the rotor of which carries no windings, the mechanical structure becomes very rugged. Operating speeds will not be high enough to cause excessive windage and friction losses. By the exercise of an ordinary amount of ingenuity, questions of slot size for windings and dimensions of pole projections on the rotating field are soluble If now, means are available for raising the frequency so generated to a value suitable for efficient radiation, we obtain another method for attacking the problem.

The static frequency doubler serves such a purpose. By its use the generated frequency may be stepped up as many times as desired. Since each step doubles the frequency, the required value is obtainable without undue complication.

Before taking up the operation of this scheme of frequency multiplication the theory involved will be discussed. If two windings be placed upon a core of ferromagnetic material, one excited by a direct current and the other carrying an alternating current, the value of the reactance presented the alternating current will vary with the degree of magnetization produced by the direct current winding because of the changing permeability of the magnetic core at different flux densities.

The magnetizing force produced by a coil in air is numerically equal to the field intensity.

$$H = \frac{0.4 \pi N I}{l} K$$

wherein N = turns,
I = current in amperes,
l = length of magnetic circuit,
K = constant, depending on shape of coil.

and the magnetic induction in a medium of permeability μ assuming the coil shape giving $K = 1$ will be

$$B = \frac{0.4 \pi N I}{l} \mu = H \mu.$$

Since $\phi = B A$,

and $L = N \frac{\phi}{10^8 i}$,

$$L = \frac{0.4 \pi N^2 A}{10^8 l} \mu$$

where A = cross sectional area of magnetic circuit,
ϕ = total flux,
L = inductance,

it is obvious that the inductance, L, varies as the total flux, which in turn is a function of the permeability μ. Hence the control of the value of reactance thru the direct current magnetization is had, because the same flux threads both coils.

The use of ferromagnetic material for carrying the lines of induction caused by currents of radio frequency brings up the question as to the character of the laws governing the flux density, depth of penetration of the flux, the iron losses, and whether the permeability varies with change of frequency. If theory confirms the practical aspect, it is to be presumed that the effective permeability and flux density under given conditions will vary inversely as the frequency. It has been shown that the depth of penetration of the flux in iron is

$$S = \frac{3570}{\sqrt{\lambda \mu f}}$$

wherein λ = conductivity,
f = frequency.

Experiments carried out along lines to confirm these assumptions have been performed by Mr. Alexanderson, and confirmation of the theory was fulfilled.

The losses in iron comprise those due to hysteresis and eddy currents, the values of which are given by the expressions,

$$P_h = N f V B^{1.6} \, 10^{-7} \text{ watts}$$
$$P_e = 1.645 \, V d^2 f^2 B^2 \, 10^{-11} \text{ watts}$$

wherein N and d are constants,
$V =$ volume of iron.

As the first increases directly and the latter as the square of the frequency it can be seen that the eddy current losses predominate at the higher frequencies. A decrease in hysteresis loss (Ewing) is suggested by reason of the rapid agitation of the molecules at the higher speeds of magnetic reversal. The total core losses in practice will be found to become less as the frequency is increased because the flux density for a given duty becomes less as the frequency is increased. Hence the losses are well within reason when using iron at radio frequencies. Combined with the fact that the effective permeability is still an appreciable quantity at these frequencies it is noted that use of ferromagnetic materials is perfectly feasible and is proving successful in practice.

The static frequency doubler consists essentially of two closed iron cores. Each of these carries a primary and a secondary alternating current winding and also a direct current magnetizing coil. The manner of connection is shown in Figure 1. The two primary windings p_1 and p_2 are wound in the same direction. The secondary windings s_1 and s_2 are reversed with respect to one another, as are also the direct current magnetizing windings m_1 and m_2. If an alternating current of potential E is applied to the primaries in series and at the same time the direct current coils are excited, a small flux change during the first half cycle will be produced in the transformer in which the alternating and direct current fluxes add. During the same half cycle in the

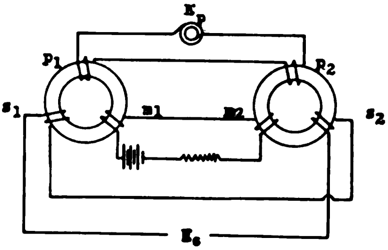

FIGURE 1

other transformer the two fluxes oppose one another and there will be a larger flux change. Since the secondary windings are reversed relative to one another, two opposing e.m.f.'s of differing amplitude will be induced in the circuit during this half cycle. Their resultant will be an e.m.f. wave having an amplitude the difference of the generated amplitudes. During the second half cycle, the phenomena will be reversed, but the resultant induced e.m.f. in the secondary circuit will be of a similar character but of course displaced in phase relative to the induced e. m. f. of the first half cycle. This results in an alternating e.m.f. at the terminals of the secondaries in series of twice the frequency of the primary e. m. f. These voltage curves are shown in Figure 2. As may be seen from this figure, the constant magnetization caused by the direct current i produces a dissymmetry in the two loops of magnetic induction, which have opposite senses with regard to the base magnetization.

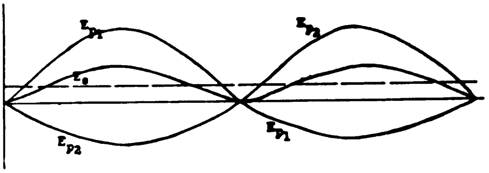

FIGURE 2

The primary voltage, e can contain only odd harmonics because of the well-known inherent characteristics of rotating generating machinery. The magnetizing current must, therefore contain even harmonics, in order to produce a flux having even harmonics, since these are not supplied by the primary e.m.f.

The action of these ferromagnetic frequency changers of the polarized core type has been given in some detail by Goldsmith.[3]

In order to bring out the theoretical treatment of these phenomena the mathematical discussions of Messrs. Kuhn[4] and Joly are here reproduced, with the notation somewhat abridged. At any time t, the resultant magnetization in core A will be

$$M_a = N I_a = N_o i_o + N i_t,$$

wherein

$N i_o$ = direct current ampere turns.
$N_o i_t$ = a. c. ampere-turns at the time t.

and the total flux in the core is

$$\Phi_a = \Phi_o + \Phi_1 \sin(\omega t - \phi_1) + \Phi_2 \sin(2\omega t - \phi_2) + \cdots$$

In core B, the resultant magnetization at any time t will be

$$M_b = N I_b = -N_o I_o + N i_t,$$

but at time t

$$= -\left\{ i_{\left(t+\frac{T}{2}\right)} N_1 - N_o I_o \right\}$$

From this we obtain the series expressing the flux in the second core

$$\Phi_b = -\Phi_o + \Phi_1 \sin(\omega t - \phi_1) - \Phi_2 \sin(2\omega t - \phi_2) + \cdots .$$

The phase angle in the second transformer is shifted one-half period with respect to the first transformer. The total primary flux Φ_p is

$$\Phi_p = \Phi_a + \Phi_b = 2\left\{\Phi_1 \sin(\omega t - \phi_1) + \Phi_3 \sin(3\omega t - \phi_3) + \cdots\right.$$

while the secondary flux is

$$\Phi_s = (\Phi_a - \Phi_b) = 2\left\{\Phi_o + \Phi_2 \sin(2\omega t - \phi_2) + \Phi_4 \sin(4\omega t - \phi_4) + \cdots .\right.$$

From this it is obvious that the secondary voltage e_2, which is generated by the flux Φ_s, contains only even harmonics.

To complete the mathematical discussion of the problem it is necessary to be able to express the $B-H$ curves of iron by an empirical formula. The necessary conditions can be embodied in an empirical expression of the following form:

$$B = A \tan^{-1} a x + C x$$

where x = ampere-turns per cm. length of magnetic circuit,
B = induction.

In order to make this function satisfy, for values of an experimentally determined curve, the three constants A, a, and C must be obtained.

From Figure 3, the three points p_1, p_2, and p_3 are chosen on the curve. These values are substituted in the above equation. A and C are eliminated as follows:—

$$B_1 = A \tan^{-1} a x_1 + C x_1$$

$$B_2 = A \tan^{-1} a x_2 + C x_2$$

$$B_3 = A \tan^{-1} a x_3 + C x_3$$

$$\frac{B_1 x_2 - B_2 x_1}{B_3 x_2 - B_2 x_3} = \frac{x_2 \tan^{-1} a x_1 - x_1 \tan^{-1} a x_2}{x_2 \tan^{-1} a x_3 - x_3 \tan^{-1} a x_2}$$

If a is obtained by trial, and substituted in two of the original equations, two linear equations in A and C result, which are easily solved. The agreement between the actual and theoretical curves is found to be very close, tho for more accurate work an equation having four arbitrary constants could be used.

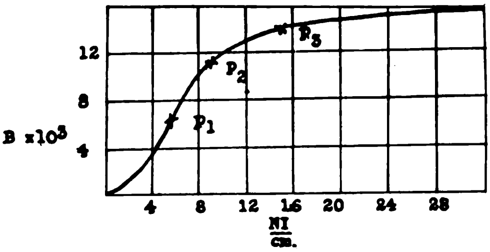

FIGURE 3

Suppose a potential e_1 is impressed on the primaries of the device. The instantaneous value of the alternating current is

$$i_1 = I \cos \omega t$$

The magnetizing ampere-turns per unit of magnetic circuit length can be written as follows:—

$$\frac{N i_1}{\text{cm.}} = x_1 \cos \omega t$$

If sufficient sized inductances are put into the direct current circuit, its reactance may be made large enough to prevent the flow of induced alternating current, and the exciting current i will contain practically no alternating current component, and therefore may be expressed as

$$i_o = I_o = \text{constant},$$

and

$$\frac{N i_o}{\text{cm.}} = x_o = \text{constant}.$$

The question now arises as to what are the forms of the secondary voltages e_a and e_b. The iron losses are neglected for the sake of simplicity. Using the relation.

$$B = A \tan^{-1} a x + C x_1 \qquad \text{where } x = \frac{N i}{\text{cm.}},$$

we can write for the flux density in the two transformers A and B:—

$$B_a = A \tan^{-1} a \left(x_1 \cos \omega t + x_o\right) + C \left(x_1 \cos \omega t + x_o\right).$$
$$B_b = A \tan^{-1} a \left(x_1 \cos \omega t - x_o\right) + C \left(x_1 \cos \omega t - x_o\right).$$

The induced emf. for an iron cross section of F sq. cm., is, for the two cases:

$$\frac{e_a}{N} = -F\frac{d\,B_a}{d\,t}\,10^{-8} \text{ (volts per turn)}$$

$$\frac{e_b}{N} = -F\frac{d\,B_b}{d\,t}\,10^{-8} \text{ (volts per turn)}$$

Substituting B_a and B_b from the previous equations we get:—

$$\frac{e_a}{N} = (F\,\omega\,10^{-8})\,x_1 \sin\omega t\left(\frac{A\,\alpha}{1+\alpha^2(x_1\cos\omega t+x_o)^2}+C\right)$$

$$= (F\,\omega\,10^{-8})\sin\omega t$$

$$\left\{\frac{A}{\alpha\,x_1}\cdot\frac{2}{\cos 2\,\omega t+4\frac{x_o}{x_1}\cos\omega t+2\left(\frac{x_o}{x_1}\right)^2+2\left(\frac{1}{\alpha\,x_1}\right)^2+1}+C\,x_1\right\}$$

$$\frac{e_b}{N} = (F\omega 10^{-8})\,x_1 \sin\omega t\left(\frac{A\,\alpha}{1+\alpha^2(x_1\cos\omega t+x_o)^2}+C\right)$$

$$= (F\,\omega\,10^{-8})\sin\omega t$$

$$\left\{\frac{A}{\alpha\,x_1}\cdot\frac{2}{\cos 2\,\omega t+4\frac{x_o}{x_1}\cos\omega t+2\left(\frac{x_o}{x_1}\right)^2+2\left(\frac{1}{\alpha\,x_1}\right)^2+1}+C\,x_1\right\}$$

These two equations differ only in a phase angle of one-half period. Hence we may make the following substitutions

$$\sin\omega t \text{ for } \sin\omega\left(t+\frac{T}{2}\right)$$

$$\cos\omega t \text{ for } \cos\omega\left(t+\frac{T}{2}\right)$$

$$\cos 2\,\omega t \text{ for } \cos 2\,\omega\left(t+\frac{T}{2}\right)$$

and the former equations may be rewritten as follows:—

$$e_b = -(F\,\omega\,10^{-8})\sin\omega\left(t+\frac{T}{2}\right)$$

$$\left\{\frac{A}{\alpha\,x_1}\cdot\frac{2}{\cos 2\omega\left(t+\frac{T}{2}\right)+4\frac{x_o}{x_1}\cos\omega\left(t+\frac{T}{2}\right)+2\left(\frac{x_o}{x_1}\right)^2+2\left(\frac{1}{\alpha x_1}\right)^2+1}+C x_1\right\}$$

also $$e_{b\,(t)} = -e_{a\left(t+\frac{T}{2}\right)}$$

On the primary side the voltages add:

$$e_1 = e_a + e_b.$$

The induced total voltage can therefore only contain odd harmonics of twice the amplitude of each voltage and in phase, since on the secondary side, where the windings are connected in opposition, all the odd harmonics disappear.

A numerical case illustrating these relations is cited. Use is made of the saturation curve of Figure 4 which was found to be represented by the equation

$$B = 9800\, tan^{-1}\left(0.41 \frac{N\, i}{\text{cm.}}\right) + 21.0 \frac{N\, i}{\text{cm.}}$$

where $A = 9.800$,
$a = 0.41$,
$C = 21.0$.

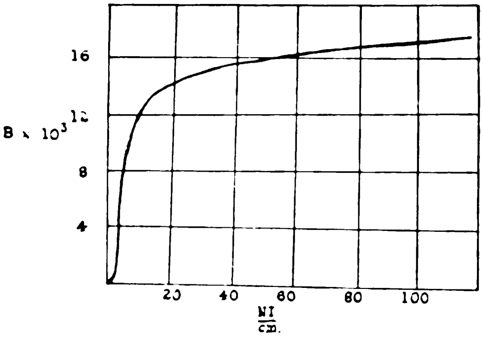

FIGURE 4

The two transformers are assumed to be saturated to the knee of the curve by means of alternating current i_1

$$x_1 = 12.2 \frac{N\, i}{\text{cm.}} . \qquad a\, x_1 = 5.0$$

Three cases are worked out with the direct current excitation having the following values

$$\frac{x_a}{x_1} = 0.25,\ 0.50,\ \text{and}\ 0.75.$$

For these values of $\frac{x_a}{x_1}$ and x_1, the curves are shown in Figures 5, 6, and 7. Upon substitution the values of e_a becomes

$\frac{x_a}{x_1} = 0.25$ . . . . $e_a{}' = \kappa\, e_a{}'$
$= 0.50$ . . . . $e_a{}'' = \kappa\, e_a{}''$
$= 0.75$ . . . . $e_a{}''' = \kappa\, e_a{}'''$

The character of each voltage curve for one core is the same for all cases, consisting of a pointed wave which is symmetrical to the right and left of the zero point, and which differs with various magnetizing excitation currents x_o only in height and phase position of the maximum ordinate.

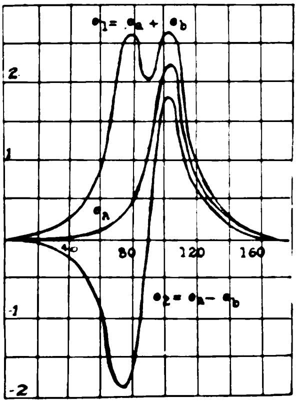

FIGURE 5

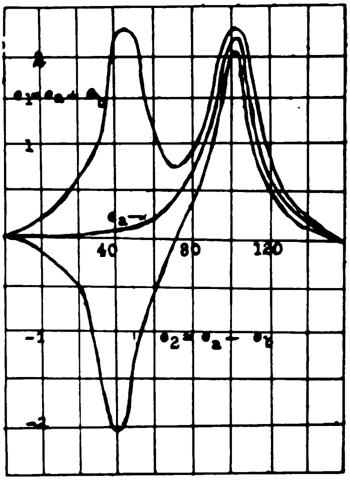

FIGURE 6

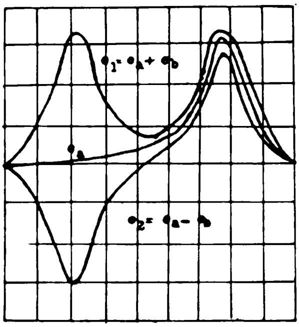

FIGURE 7

The position of the maximum ordinate can be found with sufficient accuracy by

$$\phi_{max} = -\cos^{-1}\frac{x_o}{x_1}$$

and the following values pertain to the cases given

where $$\frac{x_o}{x_1} = \begin{cases} 0.25 \\ 0.50 \\ 0.75 \end{cases} \qquad \phi = \begin{cases} 104^\circ \\ 120^\circ \\ 138^\circ \end{cases}$$

The even harmonics are related as follows:

$$e_2 = e_a - e_b = \kappa\, e_2$$

Odd harmonics are eliminated by the method of connection, that is

$$e_1 = e_a + e_b = \kappa\, e_1$$

The curves of the even harmonics are shown in Figures 5, 6, and 7.

Analysis of the total voltage to the sixth harmonic indicate that not only a strong second but also other higher harmonics exist under the condition that the magnetizing current is of sinusoidal shape.

Returning to the practical aspect concerning the actual operating conditions. A medium frequency alternator connected to a first doubling stage will have a rather high synchronous impedance and hence the transformer would only furnish a small magnetizing current. To overcome this, use is made of neutralizing capacity in this circuit and the advantages of resonance obtained. It has been demonstrated that it is possible to operate alternators of the required type in parallel, requirements for synchronizing being the same as for ordinary generators, the relation between reactance and resistance indicating the existence of certain limits within which stable operation is obtained. Therefore, if any difficulty is experienced in obtaining very large outputs from a single unit, more may be used.

Taking the hypothetical case of a required communication circuit over a distance of 1,800 miles (3,000 kilometers), the wave length for best operating efficiency is found in practice to be about 8,000 meters corresponding to approximately 38,000 cycles. If an initial generated frequency of 9,500 cycles is used, the desired operating frequency can be reached thru two doubling stages. This is indeed a simple process and with so few stages of transformation the losses will not become large and a transmitter of relatively high efficiency is available. Small desired frequency changes are obtainable thru variation of alternator speed and larger steps are attained thru fewer or more doubling stages.

Energy control with this system is facilitated by the permeability—inductance relation already discussed. Some of the possible methods have already been mentioned, and are further discussed for the sake of completeness.

If resonance is obtained by the use of series capacity in each doubler circuit, the conditions for maximum energy in the circuits will be very critical. If, now, any of the direct current magnetising values are varied, a change of reactance will take place in the alternating current circuits and hence a small variation in this circuit will produce large energy changes in the radio frequency circuits thru detuning.

Figure 8[5] illustrates a method for transmitting telegraphic signals. A key, *k*, is arranged in shunt to a resistance *R* in the auxiliary circuit, the electrical constants being so arranged that short-circuiting of this resistance will produce the necessary change of inductance in the energy circuit to disturb markedly the resonant condition.

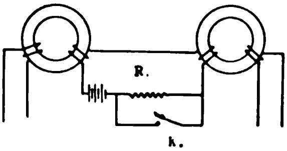

FIGURE 8

Figure 9 depicts an arrangement whereby a musical note of audio frequency may be superimposed on the radio frequency current. This is accomplished by additional coils on the cores, which coils are supplied with the required audio frequency current, thus serving to produce periodical variations in the magnetic flux.

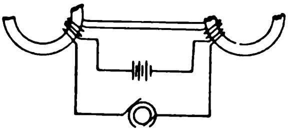

FIGURE 9

In Figure 10 is shown a method of telephonic control by means of a microphone in the auxiliary circuit.

These few illustrations of control methods show the ease with which energy manipulation is obtained in using such a system.

In order to obtain a maximum of efficiency in the antenna, the last coil of the doubling system may be connected, if desired, to an ordinary transformer, of proper dimensions, for the purpose of raising or lowering the voltage to the proper value to fit the constants of the radiating system.

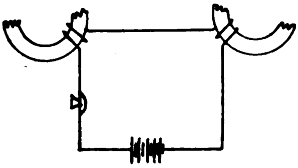

FIGURE 10

A schematic arrangement of a complete transmitter is shown in Figure 11.

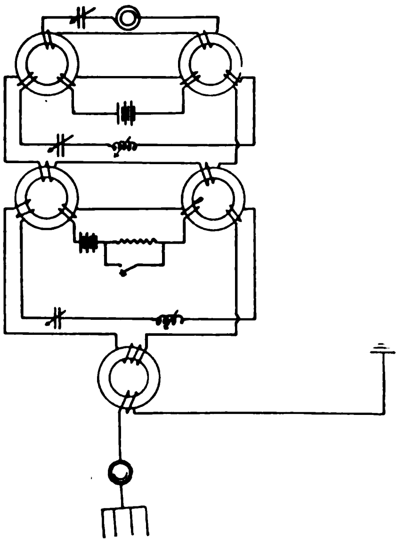

FIGURE 11

Sufficient information is now available at radio frequencies to place the design of alternators and transformers of the type required for the operation of such a system on a practical basis.

Other obstacles such as obtaining laminated core material of proper thinness, of the order of 1 to 2 mils (0.001 to 0.002 inch or 0.02 to 0.04 mm.), and the close speed regulation of prime movers, can be met to any necessary degree.

References:

1. "Magnetic Properties of Iron at Frequencies Up to 200,000 Cycles," Alexanderson, "Proceedings A. I. E. E.," 1911.
2. "The Goldschmidt System of Radio Telegraphy," Emil E. Mayer, "Proceedings I. R. E.," Volume 2, Number 1, Page 69. 1914.
3. "Radio Frequency Changers," Alfred N. Goldsmith, "Proceedings I. R. E.," Volume 3, Number 1, page 55, 1915. See also Goldsmith, "Radio Telephony" (Wireless Press, New York, 1918), pages 182 to 203.
4. "Jahrb. der Draht. Tel.," Epstein, Kuhn, and Joly, 1915, and 1916.
5. U. S. Patent number 1,181,556, Von Arco and Meissner.

SUMMARY: The system of sustained wave generation based on the use of a radio or high audio frequency alternator and frequency multipliers is considered.

The mathematical expressions for the secondary voltages of the unloaded ferromagnetic doubler of the polarized core type are derived and explained.

THEORY OF ANTENNA RADIATION*

By

A. Press

(Assistant Professor of Electrical Engineering, University of California, Berkeley, California)

The problem of antenna radiation has been attacked mathematically, notably by Hertz, Macdonald, Pearson and Lee, Love, Abraham, Bennett, and Hack. In no case, however, has the mathematical investigation included the condition of the substantially sinusoidal distribution of voltage and current along the antenna. In a series of papers published in the Proceedings of the Institute of Radio Engineers, I have shown how the variable distribution of inductivity L (self induction coefficient), and capacity C affects such distribution along the antenna. In the subjoined paper it will be shown that Eccles' formula for radiation involving as it does an equation of the following type

$$I = \frac{a}{\sqrt{r}} \cdot \varepsilon^{-\beta r}$$

is theoretically deducible, and shown to be dependent on the substantially sinusoidal voltage distribution of the antenna. This will be done without in any manner invoking the principle of the so-called Heaviside Layer. Again it will be shown that there is theoretical ground for the Cohen "dispersion factor" $(1+Ax)$, for which latter see "The Electrician," London, February 25, 1916. "Fading" will also be explained.

The present writing, it will be found, is limited to the case of a finite vertical grounded antenna with a flat earth. Cylindrical co-ordinates will, therefore, have to be resorted to. The Heaviside-Hertz equations for the field will be set up and solved by the symbolic methods of Oliver Heaviside. However, in order more easily to investigate the problem in hand it will be found desirable to split up the investigation of the radiation phenomena into two parts. These are called respectively, "voltaic effect" and "galvanic effect." The reason for this is as follows:

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* Received by the Editor, July 9, 1919.

In Maxwell's theory of the electromagnetic medium two circuital laws are made fundamental. It is necessary to consider the radiation system as really constituting an electric circuit. Current passes up, along the antenna, thru the conductive medium of the antenna wire to flow off afterwards into the dielectric medium in much the same manner as current flows across the insulating medium from the opposite sides of a condenser. From this point of view, the antenna wire and the current in it, by setting up a variable external magnetic field (and consequent dielectric stresses), produce radiation phenomena which, altho contributing nothing from an energy standpoint according to the application made of Poynting's Theorem, yet sets up distant field components that ought to be taken into account when considering a complete radiation theory. The above type of action limited as it is by the finite type of antenna is referred to as "galvanic effect." It represents the leaky transformer effect.

The second type, having reference to the rest of the electric circuit, is designated the voltaic effect, because it is found to be dependent upon the distribution of electrical potential along the antenna conductor with respect to earth. The direct effect corresponds to the so-called electrostatic induction. It also sets its magnetic field components by its variation.

Insofar as the inductivity L and capacity C per unit of length of antenna are substantially constant except for 10 per cent. of the antenna height near the bottom of the antenna and 5 per cent. near the top, the theoretical work becomes much simplified in consequence. In any case the subjoined investigation should be of interest in view of the work of Messrs. Weagant and Taylor, mentioned in the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, of June, 1919.

DIFFERENTIAL EQUATIONS FOR VERTICAL GROUNDED ANTENNA. In order to take into account the radiation in all directions from a vertical grounded antenna on a flat earth, we shall resort to the use of cylindrical co-ordinates. The antenna will be assumed as extending upward in the z direction perpendicular to the earth's surface in which the co-ordinates x and r are taken. In this case, the only component of magnetic force H_θ possible will be that having circular symmetry about the z or antenna axis. On the other hand, the displacement current components will be D_r and D_r but with no D_z component.

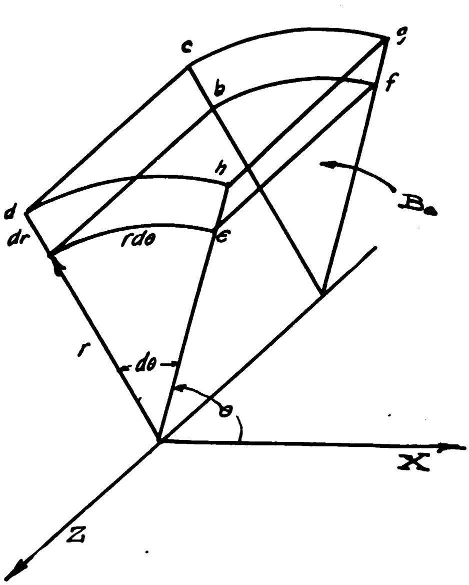

FIGURE 1

To arrive at the necessary differential equations, consider the flux passing thru the face hf. It will be $B\,dz\,dr$. Then

$$-\frac{dB}{dt}dz\,dr = \mathcal{E}_z\,dz + \left(\mathcal{E}_z + \frac{d\mathcal{E}_z}{dr}dr\right)(-dz) + \mathcal{E}_r(-dr) + \left(\mathcal{E}_r + \frac{d\mathcal{E}_r\,dz}{dz}\right)dr$$

$$\frac{dB}{dt} = \frac{d\mathcal{E}_z}{dr} - \frac{d\mathcal{E}_r}{dz}.$$

Considering now the face cf, the displacement is $D_z \cdot dr \cdot r d\theta$. We will have, therefore

$$\frac{dD_z}{dt} \cdot dr \quad r\,d\theta = H\,(-r\,d\theta) + \left(H + \frac{dH}{dr}dr\right)(r+dr)\,d\theta$$

$$\frac{dD_z}{dt} = \frac{1}{r}H + \frac{dH}{dr}.$$

Taking now the face af, the displacement is $D_r \cdot dz \cdot r\,d\theta$ Thus

$$\frac{dD_r}{dt} \cdot dz \cdot r\,d\theta = H \cdot r\,d\theta + \left(H + \frac{dH}{dz}dz\right)(-r\,d\theta)$$

$$\frac{dD_r}{dt} = -\frac{dH}{dz}$$

Introducing the relations

$$B=\mu H$$

$$D=\kappa \mathcal{E}$$

$$\mu\kappa \frac{d^2 D_r}{d t^2} = \frac{d^2 D_r}{d z^2} + \frac{d^2 D_r}{d r^2} + \frac{1}{r}\frac{d D_r}{d r} - \frac{1}{r^2} D_r.$$

Symbolizing the t and z operators, the last equation takes the form:

$$\frac{d^2 D}{d r^2} + \frac{1}{r}\frac{d D}{d r} - \left(\frac{1}{r^2}+\Delta^2\right) D = 0$$

where

$$\Delta^2 = q^2 - z_1{}^2;\quad q^2 = -\frac{1}{V}\cdot\frac{d^2}{d t^2};\quad z_1{}^2 = \frac{d^2}{d z^2}$$

and V is the velocity of light.

VOLTAIC EFFECT FROM VERTICAL GROUNDED ANTENNA. We have seen that the differential equations of condition lead to the form

$$\frac{d^2 D_r}{d r^2} + \frac{1}{r}\frac{d D_r}{d r} - \left(\frac{1}{r^2}+\Delta^2\right) D_r = 0$$

A suitable solution will be

$$D_r = r_1 \cdot K_0\,(\Delta\, r) \cdot A$$

where K_0 is a zeroth Bessel of the second kind with $r_1 = \frac{d}{d r}$ and A any desired function independent of r, but that may involve the time t or height z. For a proper solution, it will be necessary to have at

$$r = a \quad \text{that} \quad D_r = D_0$$

with a as the radius of the antenna wire and where D_0 is the known value of the displacement due to the distributed voltage v_z along the vertical aerial.

In any case, the elemental charge dQ for any height $d z$ of the aerial will be given in electrostatic units by

$$d Q = \kappa \left(\frac{d C}{d z} \cdot d z\right) v$$

if C is the capacity per unit of length. Dividing, therefore, by the elemental cylindrical area involved, $2\pi a \cdot d z$, then

$$D_0 = \frac{1}{2\pi a} \cdot \frac{d Q}{d z} = \frac{\kappa}{2\pi a} \cdot \frac{d C}{d z} \cdot v.$$

As stated above for a vertical aerial, except within ten per cent of the ground, we can take over the range $0<z<h$, where h is the antenna height

$$\frac{dC}{dz} = a$$

Moreover, for, $0<z<h$

$$v = v_z \cdot \sin pt$$

$$v_z = v_o \cdot \sin \frac{\pi z}{2h}.$$

Substituting in the above, but employing the shift operator S (see, for example, Heaviside's "Electromagnetic Theory," volume 2, page 413; volume 3, page 236), such that

$$S \cdot z^0 = \left(1 - \varepsilon^{hz_1}\right) \cdot z^0$$

$$D_r = \frac{r_1 \cdot K_o(\Delta r)}{\{r_1 K_o(\Delta r)\}_{r=a}} \cdot \frac{a\kappa}{2\pi a} \cdot v_o \cdot \sin \frac{\pi_z}{2h} \cdot S z^0 \cdot \sin pt\ .$$

from which to obtain H_θ and D_z, and so on. In reality, as will be pointed out elsewhere, we ought to take

$$v = v_z \cdot \sin(pt + \psi)$$

to allow for radiation resistance of the antenna. Radiation is assumed, in other words, to bring the voltage and current distributions into phase with one another. However, the above is sufficient to arrive at the requisite formulas.

It will be well to point out the nature of the graph belonging to

$$y = v_z \cdot S z_0 = v_o \cdot \sin \frac{\pi z}{2h} \cdot S z^0.$$

Clearly the function

$$y = S z^0$$

implies a flat top curve only over the range $0<z<h$. Therefore the function

$$y = v_o S z^0 \cdot \sin \frac{\pi z}{2h}$$

will mean that the sine function extends only over the range $o<z<h$, and that beyond, the function represents the axis of z where $y=0$. This will mean that over the range h, differentiations and integrations are to be performed on the function $v_o \cdot \sin \frac{\pi z}{2h}$ as operand; but beyond this range, the operand is zero. This distinction is very important, for integrations of zero lead to finite quantities.

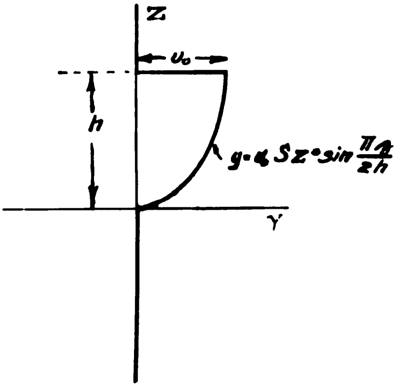

FIGURE 2

It is usual to write

$$r_1 \cdot K_o(\Delta r) = -\Delta K_1(\Delta r)$$

so that

$$D_r = \frac{K_1(\Delta r)}{K_1(\Delta a)} \cdot \frac{a \kappa v}{2 \pi a} \cdot S \cdot z^o$$

For values in which $8 \Delta r > 1$ the first term only need be taken into account in the divergent series development for $K_1(\Delta r)$ which from the generalized series

$$K_m(\Delta r) = \left(\frac{2}{\pi \Delta r}\right)^{\frac{1}{2}} \cdot \varepsilon^{-\Delta r} \left\{1 - \frac{1^2 - 4m^2}{1! \cdot 8 \Delta r} + \frac{(1^2 - 4m^2)(3^2 - 4m^2)}{2! \cdot (8 \Delta r)^2} - \right\}$$

is obtained by putting $m = 1$

$$\therefore \frac{K_1(\Delta r)}{K_1(\Delta a)} \equiv \left(\frac{a}{r}\right)^{\frac{1}{2}} \cdot \varepsilon^{-\Delta(r-a)}$$

and

$$D_r \equiv \left(\frac{a}{r}\right)^{\frac{1}{2}} \cdot \frac{a \quad v}{2 \pi a} \cdot \varepsilon^{-\Delta(r-a)} \cdot S \cdot z^o$$

Over the range $0 < z < h$, the function v occurring in D_r is a sinoidal function of z, so that within the above limits the operator involving Δ will reduce to a pure exponential. We have

$$v_z = v_o \cdot \sin \frac{\pi z}{2h}; \quad z_1^2 = -\left(\frac{\pi}{2h}\right)^2$$

moreover,

$$q^2 = -\beta^2 = -\left(\frac{2\pi}{\lambda}\right)^2$$

so that on introduction in Δ it follows

$$\Delta^2 = q^2 - z_1^2 = \left(\frac{\pi}{2h}\right)^2 - \left(\frac{2\pi}{\lambda}\right)^2$$

However, it is known that for a simple vertical antenna we can write, seemingly due to radiation resistance effects,

$$\lambda = M \cdot 4h \quad \text{with} \quad M > 1$$

That is, the wave length λ is always slightly larger than four times the vertical height h. Making the substitution for h.

$$\Delta^2 = \left(\frac{2\pi}{\lambda}\right)^2 \{M^2 - 1\} = \left(\frac{2\pi}{\lambda}\right)^2 \cdot N^2$$

$$\Delta = \frac{2\pi}{\lambda} \cdot N$$

For the range $z > h$, that is, above the antenna height, a different type of treatment will have to be resorted to.

The functions considered with $z < h$ will be primed to distinguish them from the same functions in the region $z > h$, in which latter case they will be double primed. By the formula for D_r

$$D_r' = \frac{\alpha \kappa}{2\pi\sqrt{a}} \cdot sin \frac{\pi z}{2h} \cdot sin\, pt \cdot \frac{v_0}{\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)}$$

$$i_r' = \frac{\alpha \kappa}{\sqrt{a}} \cdot V \cdot sin \frac{\pi z}{2h} \cdot cos\, pt \cdot \frac{v_0}{\lambda\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)}$$

This latter suggests very strongly the radiation formula of Eccles ("Wireless Telegraphy and Telephony," 2nd edition, page 153, London: Benn Brothers). The above formula for the radial component of current is in greater conformity with the Eccles form than with that due to Austin-Cohen.

Working backwards, since Eccles gives the coefficient

$$2\pi N = 2 \times 10^{-4}$$

$$\Delta = \frac{2 \times 10^{-4}}{\lambda}$$

It will be remembered that the condition obtaining for taking the first term of the series expansion $K_o(\Delta r)$ was that for great accuracy

$$8\Delta r > 1$$

or

$$r > \frac{1}{8} \cdot \frac{\lambda}{2 \times 10^{-4}} = \frac{10.000}{16} \lambda \equiv 600\lambda.$$

If, then, λ is in kilometers, the distance r would be in kilometers. Again from the above, we have

$$N^2 = \frac{10^{-8}}{\pi^2}; \quad M^2 = 1 + 10^{-9}$$

$$\therefore M = 1 + 5 \times 10^{-10}$$

In other words, according to the above theory, coupled with experiment for a vertical antenna, the wave length λ is very closely 4 times the antenna height.

From the componental equations, it now follows

$$H_\theta' = \frac{2a\kappa}{\pi\sqrt{a}} \cdot V \cdot \cos\frac{\pi z}{2h} \cdot \cos pt \cdot \frac{v_0 h}{\lambda\sqrt{r}} \cdot \varepsilon^{\frac{-2\pi N}{\lambda}(r-a)}$$

$$-\frac{dH_\theta'}{dt} = \frac{4a}{\mu\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdot \frac{v_0 h}{\lambda^2\sqrt{r}} \cdot \varepsilon^{\frac{-2\pi N}{\lambda}(r-a)}.$$

To find i_z we have

$$i_z = \left(\frac{1}{r} + r_1\right) H,$$

and therefore the vertical component of current density at the receiving end is

$$i_z' = \left(\frac{1}{2r} - \frac{2\pi N}{\lambda}\right) \cdot \frac{2a}{\pi V\mu\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \frac{v_0 h}{\lambda\sqrt{r}} \cdot \varepsilon^{\frac{-2\pi N}{\lambda}(r-a)} \cos pt$$

Thus, so far as the law of radiation goes, it is seen that a great deal depends on whether the reception of signals is made to depend more on i_z than on $-\frac{dH}{dt}$. In the case of i_z, the law as to distances is according to $\frac{1}{r^{\frac{3}{2}}}$; whereas as for $-\frac{dH}{dt}$, it is according to $\frac{1}{\sqrt{r}}$. So far as the vertical antenna is concerned for the voltaic effect, the greater the height h the better. The term $\left(\frac{1}{2r} - \frac{2\pi N}{\lambda}\right)$ is suggestive of the Cohen factor $(1+A\,r)$, which was suggested to account for dispersion and diffraction.

The fact that λ appears in the exponent as a denominator, as well as in the body of the equations for i_z and $-\frac{dH_\theta}{dt}$, for a vertical antenna, bespeaks an optimum wave length as Dr. Cohen first suggested.

Considering now the case where $z > h$, that is heights above the ground greater than the antenna height, the function v is zero over this range so that in D_r it is necessary to interpret the operator function

$$\varepsilon^{-\Delta(r-a)} \cdot S^1 z^0 = S \cdot \varepsilon^{-\Delta(r-a)} z^0$$

with
$$\Delta^2 = q^2 - z_1^2; \quad q = j\,p = \frac{2\pi}{\lambda} \cdot j.$$

It can be shown that the above operator function reduces to the following form:

$$\varepsilon^{-\Delta(r-a)} \cdot z'' = cos\,\frac{2\pi}{\lambda}(r-a) \cdot z'' - sin\,\frac{2\pi(r-a)}{\lambda} \cdot J_s\left(\frac{2\pi z}{\lambda}\right) \cdot j$$

where
$$J_s(\beta y) = \beta \int J_0(\beta y) \cdot d\,y.$$

Applying now the shift operator S,

$$S \cdot \varepsilon^{-\Delta(r-a)} \cdot z'' = -sin\,\frac{2\pi}{\lambda}(r-a) \cdot S J_s\left(\frac{2\pi z}{\lambda}\right) \cdot j$$

since $S z''$ is zero for $z > h$. On the other hand, $S J_s$ is given by

$$S \cdot J_{s\;z>h} = \left[J_s\left(\frac{2\pi z}{\lambda}\right) - J_s\left(\frac{2\pi(z-h)}{\lambda}\right)\right]_{z>h}$$

The formula for displacement D_r for $z > h$ now becomes instead

$$D_r'' = \frac{a\kappa}{2\pi\sqrt{a}} \cdot \frac{r_0}{\sqrt{r}} \cdot sin\,\frac{2\pi(r-a)}{\lambda} \cdot cos\,pt \cdot \left[J_s\left\{\frac{2\pi(z-h)}{\lambda}\right\} - J_s\left(\frac{2\pi z}{\lambda}\right)\right]_{z>h}$$

The attenuation factor $[\;]_{z>h}$ approaches zero for large values of h. *Thus a stationary wave is indicated so far as the voltaic effect alone is concerned, with nodal points spaced half a wave length λ apart*, that is, for the radial component of current as against i_r',

$$i_r'' = -\frac{a V\kappa}{\sqrt{a}} \cdot \frac{r_0}{\lambda\sqrt{r}} \cdot sin\,\frac{2\pi(r-a)}{\lambda} \cdot sin\,pt \cdot \left[J_s\left\{\frac{2\pi(z-h)}{\lambda}\right\} - J_s\left(\frac{2\pi z}{\lambda}\right)\right]_{z>h}$$

Again, from the auxiliary differential equations of condition—

$$H_\theta'' = -\frac{a V\kappa}{\sqrt{a}} \cdot \frac{r_0}{\lambda\sqrt{r}} \cdot sin\,\frac{2\pi(r-a)}{\lambda} \cdot sin\,pt \cdot \left[J_{s2}\left\{\frac{2\pi(z-h)}{\lambda}\right\} - J_{s2}\left(\frac{2\pi z}{\lambda}\right)\right]_{z>h}$$

if by J_{s2} is meant the integration of J_s with respect to z.

Thus, integrating—

$$\int J_s(\beta y) \cdot d\,y = \frac{(\beta y)^2}{2} - \frac{(\beta y)^4}{4 \times 3.2^2} + \frac{(\beta y)^6}{6 \times 5.2^2.4^2} - \frac{(\beta y)^8}{8 \times 7.2^2.4^2.6^2} + \cdots = J_{s2}$$

Operating, now, according to

$$i_z=\left(\frac{1}{r}+r_1\right)H_\theta$$

$$i_z''=-\left[\frac{1}{2r}+\frac{2\pi}{\lambda}\cdot cot\left\{\frac{2\pi(r-a)}{\lambda}\right\}\right]\cdot\frac{aV\kappa}{\sqrt{a}}\cdot\frac{v_0}{\lambda\sqrt{r}}\cdot sin\,\frac{2\pi(r-a)}{\lambda}\cdot sin\,pt\,\left[J_{s2}\right]_{z>h}$$

The above formulas must throw considerable light on the problem of the horizontal antenna, for the latter can be considered as made up of two abutting vertical antennas combined with an image of the whole below the surface of the earth. *In fact, "fading" as a function of the wave length changes is explained by the J_s nodal spacings.* It is therefore particularly interesting that the Eccles formula conforms to those derived for a vertical antenna proper. Moreover, so far as reception is concerned, it is seen to be important whether the reception loop be near the ground or not.

GALVANIC EFFECT FROM A VERTICAL GROUNDED ANTENNA. If the fundamental differential equations of condition are examined, it will be seen that they must obtain for galvanic effect as well. However, the conditions to be met are different in the two cases. For the present case the limiting condition is the conduction current distribution in the vertical antenna wire itself setting up a value of H_θ at the surface to correspond. It will, therefore, be necessary first to obtain the required differential equation in H_θ and then solve the same subject to the conditions just mentioned. From this latter, the values of D_r and D_z can be derived.

Insofar as

$$t_1\cdot D_r=-z_1H$$

$$H=-\frac{t_1}{z_1}\cdot D_r.$$

Then, substituting, we have

$$\frac{d^2H_\theta}{dr^2}+\frac{d^2H_\theta}{dz^2}+\frac{1}{r}\cdot\frac{dH_\theta}{dr}-\left(\frac{1}{r^2}+\frac{1}{V^2}\cdot\frac{d^2}{dt^2}\right)H_\theta=0.$$

Rewriting in symbolical notation (dropping subscripts for the time being)

$$\frac{d^2H}{dr^2}+\frac{1}{r}\frac{dH}{dr}-\left(\frac{1}{r^2}+\Delta^2\right)H=0$$

where $$\Delta^2=q^2-z_1^2$$

We have seen that a solution will be of the form

$$H = \frac{K_1(\Delta r)}{K_1(\Delta a)} \cdot A$$

where A is again any desired function, but independent of r, but that may involve the time t and the height z. Evidently where $r = a$

$$H_{r=a} = H_o = A$$

and must accord with the distribution of H at the surface of the antenna wire. Because, in electromagnetic units,

$$H_o \times 2\pi a = I$$

$$A = \frac{I}{2\pi a}$$

However, the current I in the conductor being distributed, will conform to a function of the following type:

$$I = I_z \cdot \cos\frac{\pi z}{2h} \cdot S z^o$$

with $$I_z = I_o \cdot \sin pt,$$

and the shift operator function $S z_o$, implying that no current I occurs where $z > h$. Hence the formula for H_θ is given by

$$I_\theta = \frac{K_1(\Delta r)}{K_1(\Delta a)} \cdot \frac{I_o}{2\pi a} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdot S z^o$$

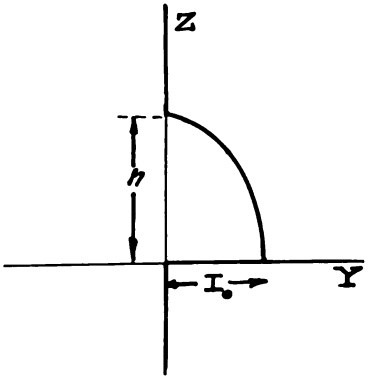

Figure 3

Again here it is necessary to remember that the sinoid $\cos\frac{\pi z}{2h}$

has two zero branches beyond the range $0<z<h$, so that H_θ will have two sets of solutions. In the range h, the solution will be primed as before, and in the region $z>h$, double primes will be employed. Considering the h region first, then, as before,

$$\Delta = \frac{2\pi N}{\lambda}$$

$$H_\theta' = \frac{1}{2\pi\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdot \frac{I_o}{\sqrt{r}} \cdot \varepsilon^{\frac{-2\pi N}{\lambda}(r-a)} \cdot Sz''$$

$$-\frac{dH_\theta'}{dt} = \frac{V}{\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdot \frac{I_o}{\lambda\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)} \cdot S \cdot z^o$$

To obtain D_r', we note first

$$i_r = -z_1 H$$

$$i_r = \frac{1}{4} \sin\frac{\pi z}{2h} \cdot \sin pt \cdot \frac{I_o}{\sqrt{a} \cdot h \cdot \sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)}$$

$$D_r' = \frac{1}{8\pi V} \cdot \sin\frac{\pi z}{2h} \cdot \cos pt \cdot \frac{I_o \lambda}{\sqrt{a} \cdot h\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\Delta}(r-a)}$$

Again, because

$$i_z = \left(\frac{1}{r} + r_1\right) H,$$

we have

$$i_z' = \left(\frac{1}{2r} - \frac{2\pi N}{\lambda}\right) \frac{1}{2\pi} \cdot \frac{1}{\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdot \frac{I_o}{\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)}$$

$$D_z' = -\left(\frac{1}{2r} - \frac{2\pi N}{\lambda}\right) \cdot \frac{1}{4\pi^2 V\sqrt{a}} \cdot \cos\frac{\pi z}{2h} \cdot \cos pt \cdot \frac{I_o \lambda}{\sqrt{r}} \cdot \varepsilon^{-\frac{2\pi N}{\lambda}(r-a)}$$

The double prime components occurring in the region $z>h$ can be obtained in a similar manner to the procedure for the voltaic radiation. The formula i_z' for the galvanic component again shows the Cohen factor tho of somewhat different form from the corresponding voltaic component.

ANTENNA RADIATION RESISTANCE OF VERTICAL GROUNDED ANTENNA. If we attempt to apply the Poynting radiation theorem to a vertical grounded antenna, it will be sufficient to know the values of $\mathcal{E}$ and H over the surface of the aerial for finding the instantaneous watts radiated. Taking the current on the antenna as given by

$$I_z = I_o \cdot \sin pt$$

it will be necessary to take

$$v_z = v_0 \cdot \sin\,(pt+\psi)$$

as against

$$v_z = v_0 \cos pt$$

which would be the case of a non-radiating transmission line. (In reality, ψ is a slowly varying function of z). Since

$$v_z = v_0 \left\{ \cos\psi \cdot \sin pt + \sin\psi \cdot \cos pt \right\}$$

It will be necessary in the formula developed involving $v = v_0 \sin pt$ to replace pt by $(pt+\psi)$ in order to allow for the wattage loss by radiation. At the surface of the aerial wire $(r=a)$, the voltaic displacement D_r is given, and therefore

$$\mathscr{E}_r = \frac{1}{\kappa} D_r = \frac{a}{2\pi a} \cdot \sin\frac{\pi z}{2h} \cdot v_0 \cdot \sin(pt+\psi) \cdots (\text{voltaic, } r=a) \quad (1)$$

The value of H_θ therefore reduces to

$$H_\theta = \frac{2a\kappa}{\pi a} \cdot V \cdot \cos\frac{\pi z}{2h} \cos\,(pt+\psi) \cdot \frac{v_0 h}{\lambda} \cdots (\text{voltaic}, r=a) \quad (2)$$

Again

$$E_z = \frac{a h v_0}{\pi^2 a}\left(\frac{1}{2a} - \frac{2\pi N}{\lambda}\right) \cdot \cos\frac{\pi z}{2h} \cdot \sin\,(pt+\psi) \cdots (\text{voltaic, } r=a) \quad (3)$$

Taking now the galvanic components by

$$H_\theta = \frac{I_0}{2\pi a} \cdot \cos\frac{\pi z}{2h} \cdot \sin pt \cdots (\text{galvanic, } r=a) \quad (4)$$

$$\mathscr{E}_r = \frac{1}{\kappa} D_r = -\frac{I_0}{8\pi a V \kappa} \cdot \frac{\lambda}{h} \cdot \sin\frac{\pi z}{2h} \cdot \cos pt \quad (5)$$

$$E_z = \frac{1}{\kappa} D_z = -\frac{I_0 \lambda}{4\pi^2 V \kappa a} \cdot \left(\frac{1}{2a} - \frac{2\pi N}{\lambda}\right) \cdot \cos\frac{\pi z}{2h} \cos pt \quad (6)$$

Considering the time elements separately in the radiation formula with vector notation $div\,\mathbf{W} = div\ V\mathbf{E} \cdot \mathbf{H}$, it is necessary to note that any time components such as $\sin pt \cdot \cos pt$. must reduce to zero over a cycle period $T = \frac{1}{f}$. The true wattage components in the formula necessitates that $\frac{d\mathscr{E}}{dt}$ give an average resultant value when multiplied by $\frac{dH}{dt}$. Combining therefore (4) with $\sin pt. \cdot$ elements in (1), (2), and (3), let

$$\frac{I_0}{2\pi a} = A; \quad \frac{2a\kappa}{\pi a} \cdot V \cdot \frac{v_0 h}{\lambda} = M$$

$$\frac{I_0}{8\pi a V\kappa} \cdot \frac{\lambda}{h} = B; \quad \frac{I_0}{4\pi^2 V\kappa} \cdot \frac{\lambda}{a}\left(\frac{1}{2a} - \frac{2\pi N}{\lambda}\right) = C$$

Then for the $\sin pt$ terms only

$$H_s \cdot \sin pt = \left\{A \cdot \cos\frac{\pi z}{2h} - M \sin\psi \cdot \cos\frac{\pi z}{2h}\right\} \sin pt =$$

$$(A - M \sin\psi) \cdot \cos\frac{\pi z}{2h} \cdot \sin pt.$$

Again, let

$$\frac{v_0 a}{2\pi a} = N; \quad \frac{a h v_0}{\pi^2 a}\left(\frac{1}{2a} - \frac{2\pi N}{\lambda}\right) = P$$

For the $\sin pt$ electric components together

$$\mathcal{E}_s \cdot \sin pt. = \left\{N \cdot \sin\frac{\pi z}{2h} + P \cdot \cos\frac{\pi z}{2h}\right\} \cos\psi \cdot \sin pt.$$

For the $\cos pt$ terms in H,

$$H_c \cdot \cos pt = M \cos\left(\frac{\pi z}{2h}\right) \cdot \cos\psi \cdot \cos pt$$

On the other hand, for the electric terms combined:

$$\mathcal{E}_c \cdot \cos pt = \left\{N \cdot \sin\psi \cdot \sin\frac{\pi z}{2h} - B \cdot \sin\frac{\pi z}{2h} - C \cos\frac{\pi z}{2h}\right\} \cdot \cos pt$$

$$= \left\{(N \sin\psi - B) \sin\frac{\pi z}{2h} - C \cos\frac{\pi z}{2h}\right\} \cos pt$$

Over an element of area $2\pi a \cdot dz = dS$ with forces $\mathcal{E}$ and H, the average wattage will be

$$dW = 2\pi a \cdot dz \, \frac{1}{T}\int_0^T (\mathcal{E}_c \cos pt + E_s \sin pt)\,(H_c \cos pt + H_s \sin pt)\,dt$$

$$= 2\pi a \cdot dz \cdot \frac{1}{T}\int_0^T (\mathcal{E}_c H_c \cos^2 pt + E_s H_s \sin^2 pt)\,dt.$$

Since over a period $T = \frac{1}{f}$, terms involving $\sin pt \cdot \cos pt$ reduce to zero. We can therefore investigate the terms separately

$$dW_c = \pi a \cdot dz \cdot \mathcal{E}_c \cdot H_c; \quad dW_s = \pi a \cdot dz \cdot \mathcal{E}_s H_s.$$

However,

$$\mathcal{E}_c H_c = \left\{(N \cdot \sin\psi - B) \sin\frac{\pi z}{2h} - C \cos\frac{\pi z}{2h}\right\} M \cos\frac{\pi z}{2h} \cdot \cos\psi$$

$$= MN \cdot \sin 2\psi \cdot \sin\frac{\pi z}{2h} - \frac{MB}{2} \cos\psi \cdot \sin\frac{\pi z}{2h} - MC \cdot \cos\psi \cdot \cos^2\frac{\pi z}{2h}.$$

Similarly, therefore

$$\mathcal{E}_s H_s = \frac{A\,N}{2} \cdot sin\frac{\pi z}{2h} \cdot cos\psi - \frac{MN}{4} \cdot sin\,2\psi \cdot sin\frac{\pi z}{2h} +$$
$$AP\,cos\psi \cdot cos^2\frac{\pi z}{2h} - \frac{MP}{2} sin\,2\psi \cdot cos^2\frac{\pi z}{2h}.$$

Adding the above terms together we note

$$A\,P = M\,C; \quad M\,B = A\,N$$

$$M\,P = \frac{2\,a^2\,V\kappa}{\pi^3} \cdot \frac{h^2}{\lambda\,a^2} \cdot v_o^2 \left(\frac{1}{2\,a} - \frac{2\,\pi\,N}{\lambda} \right)$$

$$\therefore d\,W = d\,W_s + d\,W_c = \frac{a^2\,V\kappa}{\pi^2} \cdot \frac{h^2}{a\,\lambda} \cdot v_o^2 \left\{ \frac{1}{2\,a} - \frac{2\pi\,N}{\lambda} \right\} sin\,2\psi \cdot cos^2\frac{\pi z}{2h} \cdot d\,z\,.$$

Integrating over the whole length of the aerial, then, because

$$\int_o^h cos^2\frac{\pi z}{2h} \cdot d\,z = \frac{h}{2}$$

$$W_h = \frac{a^2\,V\kappa}{\pi^2} \cdot \frac{h^3}{2\,\lambda\,a} \cdot v_o^2 \left\{ \frac{1}{2\,a} - \frac{2\,\pi\,N}{\lambda} \right\} sin\,2\psi$$
$$\frac{a^2\,V\kappa}{2\,\pi^2} \cdot \frac{h^3}{\lambda\,a^2} \cdot v_o^2 \cdot sin\,\psi \cdot cos\,\psi.$$

since N is small. The above represents the radiation from the body of the wire only and does not include in any manner the radiation by means and thru the earth's surface.

To evaluate ψ, the wattage can be derived from a consideration of the current and voltage in the aerial. The average wattage consumed in the antenna at any distance z and over the interval $d\,z$ is given in terms of maximum voltage and current by

$$d W_h = \frac{d\,z}{T} \int_o^T \left\{ I_o \cdot cos\frac{\pi z}{2h} \cdot sin\,pt \right\} \frac{d}{dz} \left\{ v_o \cdot sin\frac{\pi z}{2h} \cdot sin(pt + \psi) \right\} \cdot dt$$
$$= \frac{d\,z}{T} \int_o^T I_o\,v_o \cdot \frac{\pi}{2\,h} \cdot \left(cos\frac{\pi z}{2h} \right)^2 \cdot (sin\,pt)^2 \cdot cos\,\psi \cdot dt,$$

because we can neglect the $sin\,pt \cdot cos\,pt$ term over an entire interval T. The above reduces to

$$d\,W_h = \frac{\pi}{4\,h} \cdot I_o\,v_o \cdot cos\,\psi \cdot \left(cos\frac{\pi z}{2h} \right)^2 \cdot d\,z$$

Integrating, therefore, over the whole length h of the aerial:

$$W_h = \frac{\pi}{8} \cdot I_o\,v_o \cdot cos\,\psi$$

This latter must equate to the value given previously. Thus

$$\frac{\pi}{8} I_o r_o \cos\psi = \frac{a^2}{2\pi^2 \sqrt{\mu}} \cdot \frac{h^3}{\lambda a^2} \cdot r_o^2 \cdot \sin\psi \cos\psi$$

$$\sin\psi = \frac{\pi^3}{4 a^2} \cdot \frac{\lambda a^2}{h^3} \cdot \sqrt{\mu} \frac{I_o}{r_o}$$

To express the above in terms of the radiation resistance, let this latter be defined by

$$\frac{W_h}{(I_o \sqrt{2})^2} = \frac{2W}{I_o^2} = R_h = \frac{\pi}{4} \cdot \frac{I_o r_o \cdot \cos\psi}{I_o^2} = \frac{\pi}{4} \cdot \cos\psi \cdot \frac{r_o}{I_o}.$$

For any vertical type of aerial the ratio $\frac{r_o}{I_o}$ is a known constant, and depends merely on the ratio of the inductance to the permittance per unit of length. Defining this ratio by

$$\Omega = \frac{r_o}{I_o}; \quad R_h = \frac{\pi}{4} \cdot \Omega \cdot \cos\psi$$

For convenience, let

$$A = \frac{\pi^3}{4 a^2} \cdot \frac{\lambda a^2}{h^3},$$

then

$$\sin\psi = A \frac{\sqrt{\mu}}{\Omega}; \quad \cos\psi = \sqrt{1 - \left(\frac{A\sqrt{\mu}}{\Omega}\right)^2}$$

$$\therefore R_h = \frac{\pi}{4} \Omega \sqrt{1 - \left(\frac{A\sqrt{\mu}}{\Omega}\right)^2} = \frac{\pi}{4} \sqrt{\Omega^2 - \left\{\frac{\pi^3}{4 a^2} \cdot \frac{\lambda a^2}{h^3} \cdot \sqrt{\mu}\right\}^2}$$

Thus, for radiation from the antenna wire itself in the case of a vertical aerial, the height should be as great as possible; for since the wave length is proportional to the height, an increase in height will more than compensate for the increase in wave length. Observe that the above takes no note of radiation from the surface of the earth. The constant a is defined by the relation $\frac{dC}{dz}$ above. For its evaluation, see the paper by the author on "The Vertical Grounded Antenna as a Generalized Bessels Antenna," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, December, 1918, volume 6, number 6.

SUMMARY: After considering the previous investigations of antenna radiation, the author attacks this problem taking into account the sinusoidal distribution of current and voltage along the antenna. The vertical grounded antenna only is considered.

A detailed mathematical investigation of radiation and field strengths at remote points follows. Formulas resembling the Eccles radiation formula, and the Cohen dispersion and diffraction factor formula are obtained and interpreted.

VOLUME 8 FEBRUARY, 1920 NUMBER 1

B4

PROCEEDINGS
of
The Institute of Radio Engineers

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
THE INSTITUTE OF RADIO ENGINEERS
THE COLLEGE OF THE CITY OF NEW YORK
140th Street and Convent Avenue, New York, N. Y.

Subscription $9.00 per Annum in the United States
$9.60 in all other Countries

GENERAL INFORMATION AND SUBSCRIPTION RATES ON PAGE 1

Entered as second-class matter, February 16, 1916, at the Post-Office, at New York, N. Y., under the Act of March 3, 1879.

Acceptance for mailing at special rate of postage provided for in Section 1103, Act of October 3, 1917, authorized October 7, 1918.

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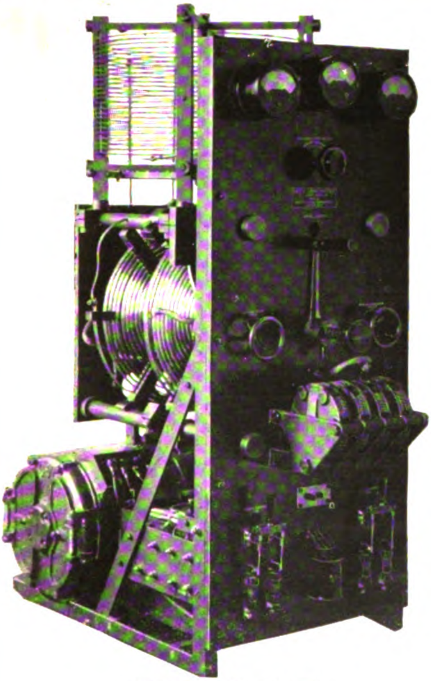

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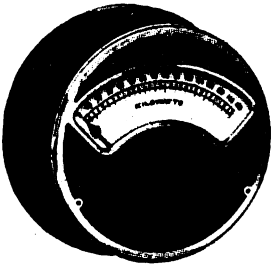

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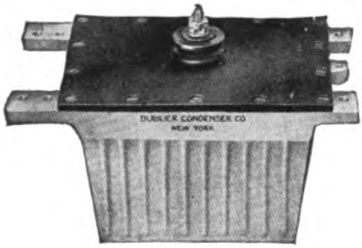

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The Research Department of the Radio Corporation of America has vacancies for several qualified Radio Draftsmen. Applicants having ALL the following qualifications ONLY will be considered.

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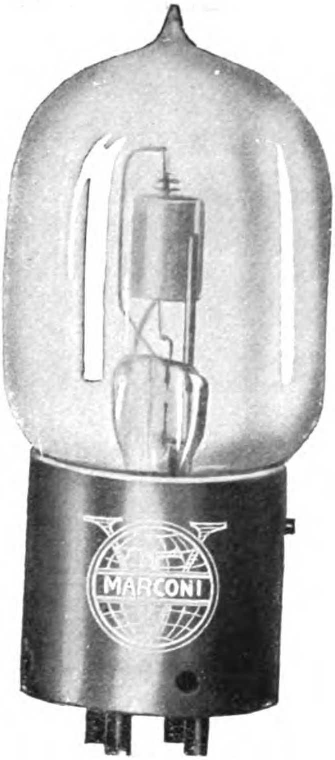

Fleming Pat. No. 803,684
De Forest Pat. Nos. 841,387-879,532

A WARNING

To Manufacturers
Importers
Dealers
Jobbers
Agents
Amateurs
Purchasers
Users of

Vacuum Tubes
The Marconi V. T. Patent is Basic

United States Letters Patent to Fleming. No. 803,684, November 7, 1905 has been held to be valid by Judge Mayer of the United States District Court for the Southern District of New York, and by the United States Circuit Court of Appeals for the Second Circuit.

It is a basic patent and controls broadly all vacuum tubes used as detectors, amplifiers or oscillions in radio work.

No one is authorized to make, sell, import or use such tubes for radio purposes, other than the owners of the patent and licensees thereunder. Any others making, selling, importing or using them alone or in combination with other devices, infringe upon the Fleming patent and are liable to a suit for injunction, damages and profits. And they will be prosecuted.

THE AUDIOTRON AND THE LIBERTY VALVE ARE NOT LICENSED UNDER THE FLEMING PATENT.

The price of the genuine Marconi V. T. delivered is $7.00 each. The standardized socket is $1.50 additional. The standard resistance, complete, costs $1.00 and is made in the following sizes: ½ megohm, 1 megohm, 2 megohms, 4 megohms, 6 megohms.

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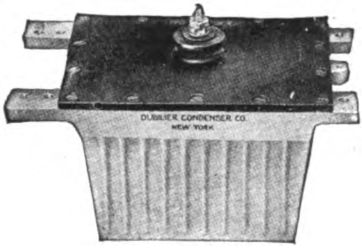

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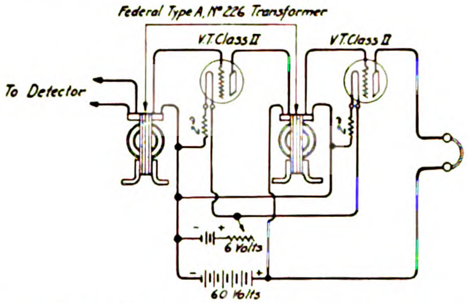
Type A



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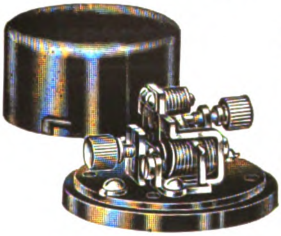

The Radio Officers of the NC Planes after testing all other buzzers, decided to use the "Mesco" on their radio equipment. The R-34 was equipped with two Mesco Radio Buzzers.

Why? Because of its reliability and constancy in operation; greater output efficiency; ease of adjustment; unaffected by extreme variations in weather conditions; exposed wires eliminated.

Sparking is almost entirely eliminated, so that the energy lost in light and heat in the operation of other buzzers is here conserved and radiated in the form of oscillating energy.

This buzzer maintains a constant note and is recommended as an exciter for checking wave-meters where pure note and ample energy are required.

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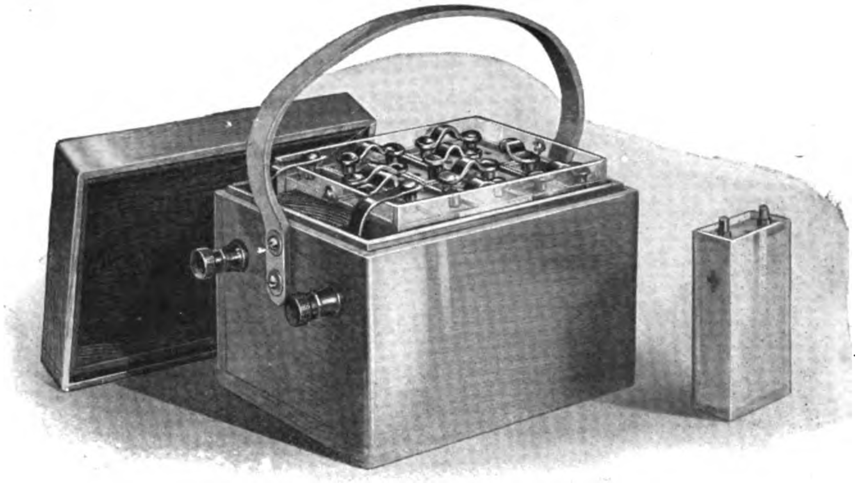

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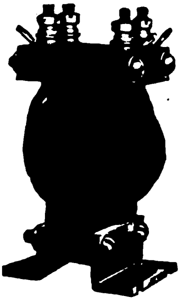

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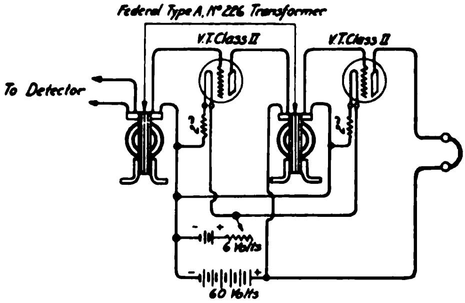

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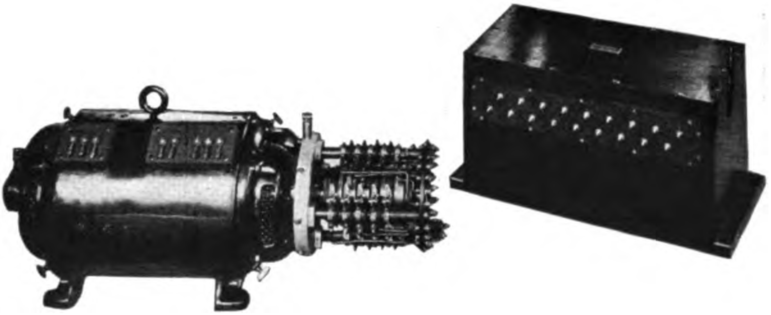

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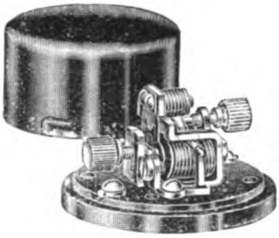

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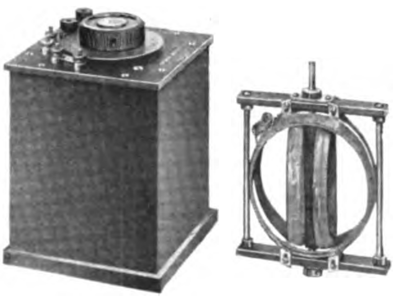

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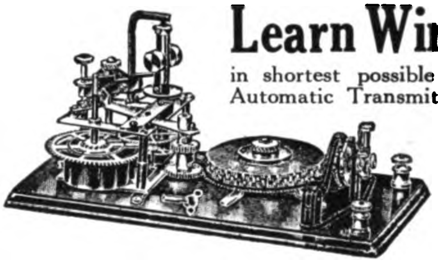

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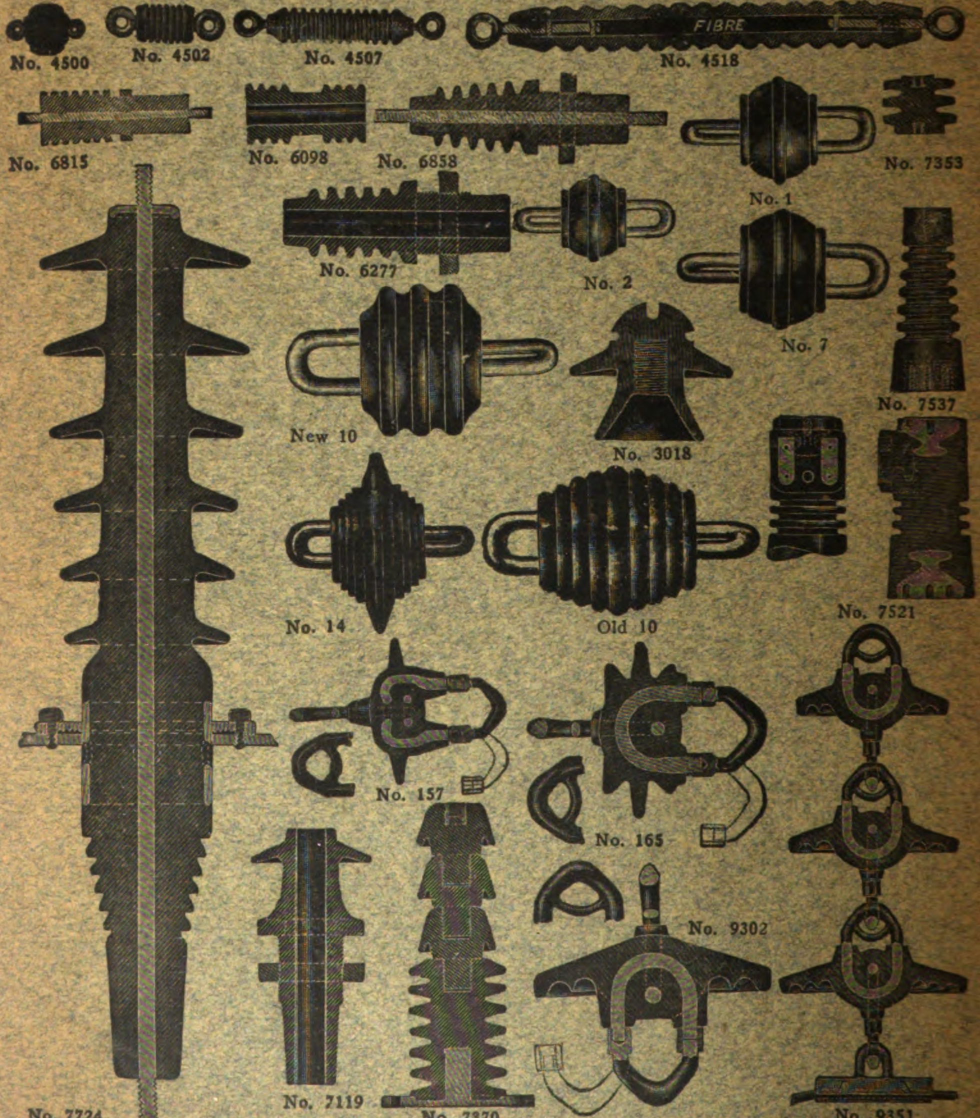

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VOLUME 8 DECEMBER, 1920 NUMBER 6

B5

PROCEEDINGS
of
The Institute of Radio Engineers

EDITED BY

ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
THE INSTITUTE OF RADIO ENGINEERS
THE COLLEGE OF THE CITY OF NEW YORK
140th Street and Convent Avenue, New York, N. Y

Subscription $9.00 per Annum in the United States
$9.60 in all other Countries

GENERAL INFORMATION AND SUBSCRIPTION RATES ON PAGE 449

Entered as second-class matter, February 16, 1916, at the Post-Office, at New York, N. Y., under the Act of March 3, 1879.

Acceptance for mailing at special rate of postage provided for in Section 1103, Act of October 3, 1917, authorized October 7, 1918.

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"Arcover" disk strain and suspension line insulators, "Safety Strain" insulators, Disk, Hood, Thimble and Rod type strain insulators, Spool and Bracket-Arm strain insulators, Pin type line insulators, Insulator pins, Roof, Barrier and Wall Bushings and Insulated Connectors, single and multi-part, plain and locking, water and gas tight, Bus-bar insulators, Insulating Supports and Pedestals, Arc Lamp Insulators, Insulating Knobs and Handles, Insulating Sheets, Tubes and Rods, etc., etc.

Electrose Railway Line Material:—"Safety Strain" Insulators, Brooklyn Strain Insulators, Insulated Turn-Buckles, Caps and Cones, "Ideal" Locking Caps and Cones, Insulated Bolts, Feed Wire Insulators, Third Rail Insulators, also special forms of Trolley Wire Suspension Insulators, etc., etc. Complete line always carried in stock.

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"Electrose" insulation (certain grades) is recognized as the best obtainable for use in connection with **insulating parts for ignition service for Automobiles, Aeroplanes, Motor Boats, etc.**

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Complete ship radio installations using damped or undamped waves. The Company will contract, at a fixed rate, for complete maintenance and operation of ship stations. Inspection and Repair Service Stations at important ports, including the facilities of the Westinghouse Electric & Manufacturing Company. Traffic Accounting and Routing.

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326 Broadway New York

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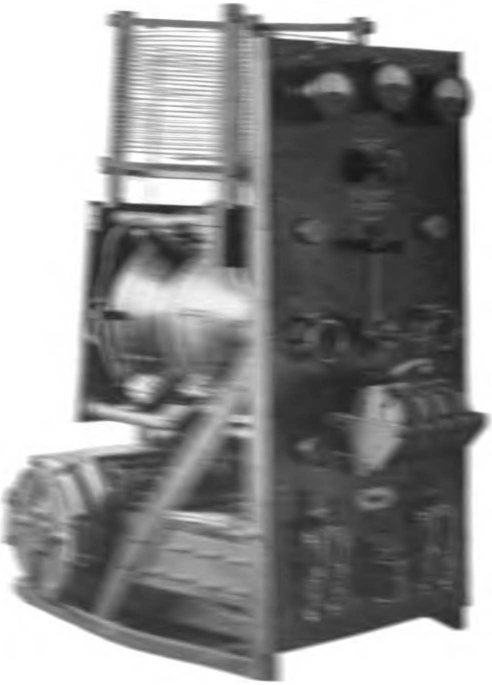

2 K.W. Transmitter

[illegible]

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FARADON

TRADE MARK

MICA CONDENSERS

[illegible]

[illegible]

TRADE MARK

TWO STEP AMPLIFIER TYPE [illegible]

[illegible]

WIRELESS SPECIALTY APPARATUS CO.

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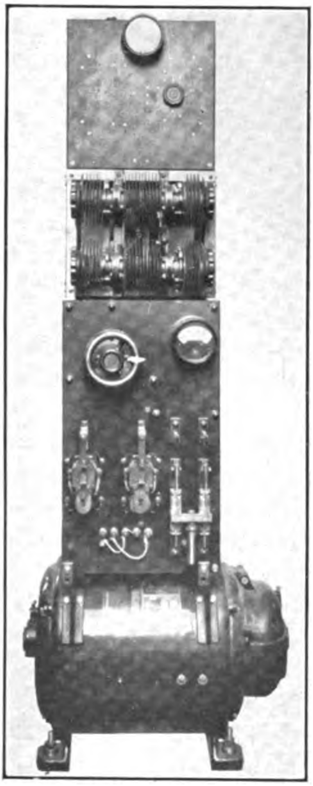

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No Critical Adjustments

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GENERAL OFFICES:

6 AND 8 WEST 48TH STREET, NEW YORK, N. Y.

Cable Address: CUTWASH, New York

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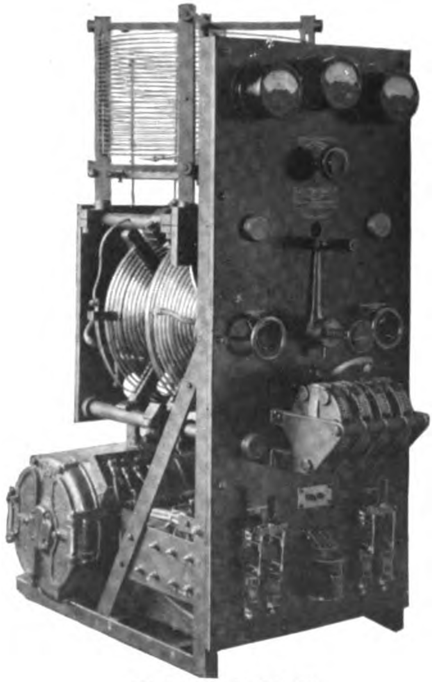
We design and manufacture radio equipment of every description from the experimental station to the complete commercial communication system.



1/2 K.W. Transmitter

Type Q.S. 500

The extreme simplicity, ruggedness and high electrical safety factor of this equipment will be quickly appreciated by the radio engineer and ship owner. This transmitter is an ideal installation for yachts, commercial vessels, colleges and research laboratories.

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Competitive tests have shown the W. S. A. **FARADON** to be the best mica primary condenser yet produced. The .004 mfd **FARADON** is tested for thirty minutes at 21,000v. and 18a. in a circuit of 1,000 meters, with a temperature rise of less than 3.5° C. Every condenser is guaranteed for one year. We also make mica protective condensers and laboratory standards.

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This amplifier is an attractive, compact unit of the resonance type, and provides a maximum of low frequency amplification.

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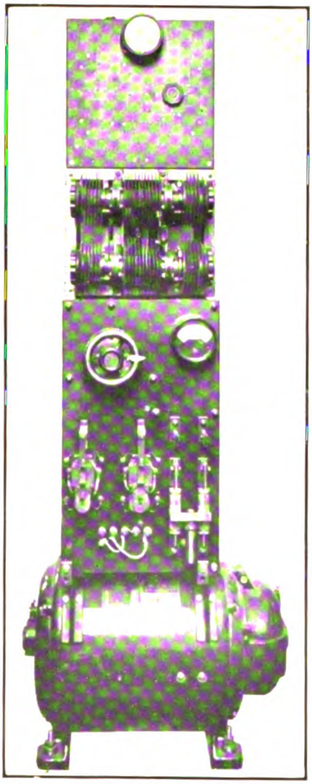

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Roller-Smith Four-Inch type W radio frequency ammeters (switchboard model) are hot wire or hot strip expansion type instruments which measure accurately and reliably currents of any radio frequency.

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Portable radio frequency instruments with the same type of mechanism for inspector's use and for laboratory use are furnished also.

Let us know your requirements and we will be pleased to submit further details.

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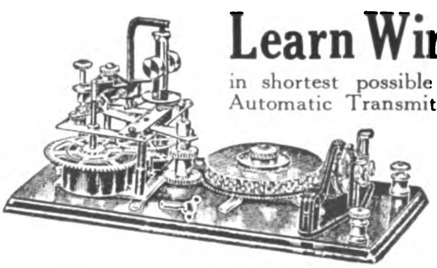

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DE FOREST

Christmas Hints for the Radio Amateur

TYPE T-200

Multi-Wave Tuner

Wave Length, 150-25,000 Meters

TYPE P-300

Combination Audion, Ultra-Audion and One-Step Amplifier

PUT DeForest Radio Apparatus on your Christmas list. Here are two instruments that will greatly increase the efficiency of your set.

TYPE T-200

Multi-Wave Tuner

Responds to any wave length—150 to 25,000 meters. Contains triple adjustable coil mountings; Vernier primary and secondary tuning condensers; primary condenser switch; all mounted on engraved Bakelite panel and in handsome cabinet, size 13 x 18 x 12¾. Price (without coils) $85.00.

TYPE P-300

Combination Audion, Ultra-Audion and One-Step Amplifier

Specially designed to fill a real need for an instrument of this kind in the average amateur station. When used in connection with the T-200 Tuner shown above, it completes a receiver of unequaled efficiency. Cabinet size 12½ x 8¾ x 7¾. Price (without tubes) $75.00.

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Inventors and Manufacturers of High Grade Radio Apparatus

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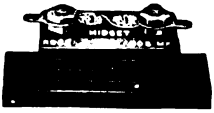
The losses in inefficient or poorly-designed condensers become very considerable and may appreciably reduce the operating efficiency of many Radio circuits. Usually the Radio man is most critical in his selection of the major parts of his apparatus, but gives very little thought to the merits of a carefully-designed fixed condenser.

The

Type ROCC .0002 mf

GREBE RADIO



Type ROCF .005 mf

Midget Fixed Condensers

are carefully designed and are constructed of mica, copper and Bakelite-Dilecto. Grebe Midget Condensers will carry a potential up to 1000 volts without danger of breakdown. The various sizes permit the use of these condensers individually or as combination capacities in almost any Radio circuit. The dimensions of the base are 1"x2"; the height varying with the capacity, from 3/8" to 7/8".

There are any number of fixed condensers on the market, but you will find only in Grebe Midget Condensers the painstaking precision that has made Grebe Radio apparatus the acknowledged standard in the amateur and experimental fields.

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NOTICE

In the infringement suit Dubilier Condenser Co., Inc., plaintiff, vs. Wireless Specialty Apparatus Co., defendant, a final decree has been handed down by his honor, Augustus N. Hand, U. S. District Judge for the Southern District of New York, holding

"That said Letters Patent Nos. 1,229,914 and 1,229,915 are good and valid in law as to the second, third, fourth, fifth, seventh, eighth, ninth, twelfth, fourteenth and fifteenth claims of Letters Patent No. 1,229,914 and as to the first, eighth, twelfth, sixteenth and seventeenth claims of Letters Patent No. 1,229,915." We hereby give notice that it is the intention of this Company to uphold to the full extent its rights under the above and all other patents held by us, and any one making, using or dealing in condensers covered by our patents without a license from us, will be held responsible to the full extent of the law.

Dubilier Condenser Co., Inc.

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Murdock Radio Receivers

An inexpensive, surprisingly sensitive, thoroughly workmanlike head set, universally conceded "the most for the money." Their record for sensitive service in thousands of stations, amateur and commercial, is their best advertisement.

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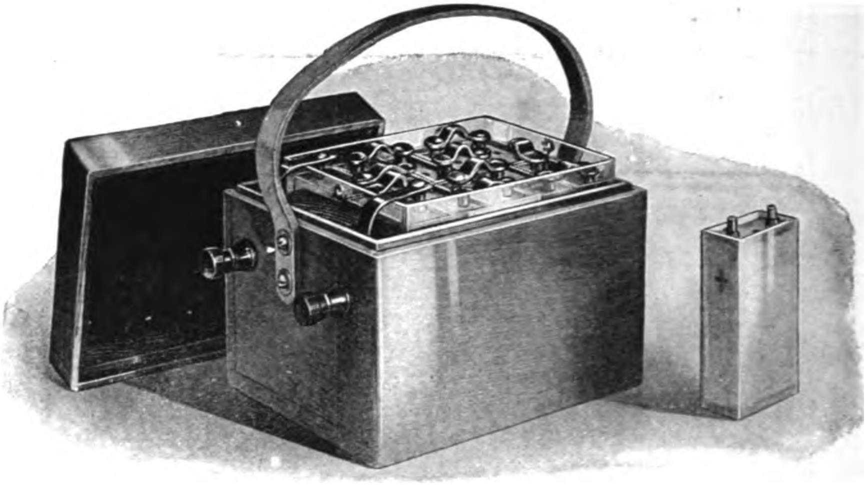

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Here is a storage battery that is shipped to you dry, may be stored indefinitely without deterioration; yet can be put in service in 6 hours.

Has a high charging rate—(0.5 amp. for 2.0 amp. unit), yet holds its voltage over long periods of continuous or intermittent work.

Delivers more energy per pound than was ever before thought possible; yet will be the strongest part of your equipment.

Comprises plates that are active ALL the way thro, but are thicker and stronger than those in batteries 50 times their size.

ACTUAL TESTS MADE IN SERVICE

No. 1—Unit fully charged—laid on shelf (no load) for 100 days. At the end of this time the open circuit voltage was 20.5. This unit was then put in service and gave 336 hours of continuous service at .005 amp. down to a closed circuit voltage of 17.2 volts.

No. 2 —Unit was fully charged—then discharged—at 0.005 amp. giving 525 hours down to 17.0 volts.

No. 3—Fully charged—discharged at 0.100 amp. down to 17.0 v. giving 22.0 hours.

No. 4—Unit was discharged at a rate which brought the voltage to 17.0 in 888 hours.

The battery was not in any way injured by these experiments, and continued to give service at its full capacity on various other rates.

The above figures are GUARANTEED. Each and
every Gem Unit will give YOU the same service

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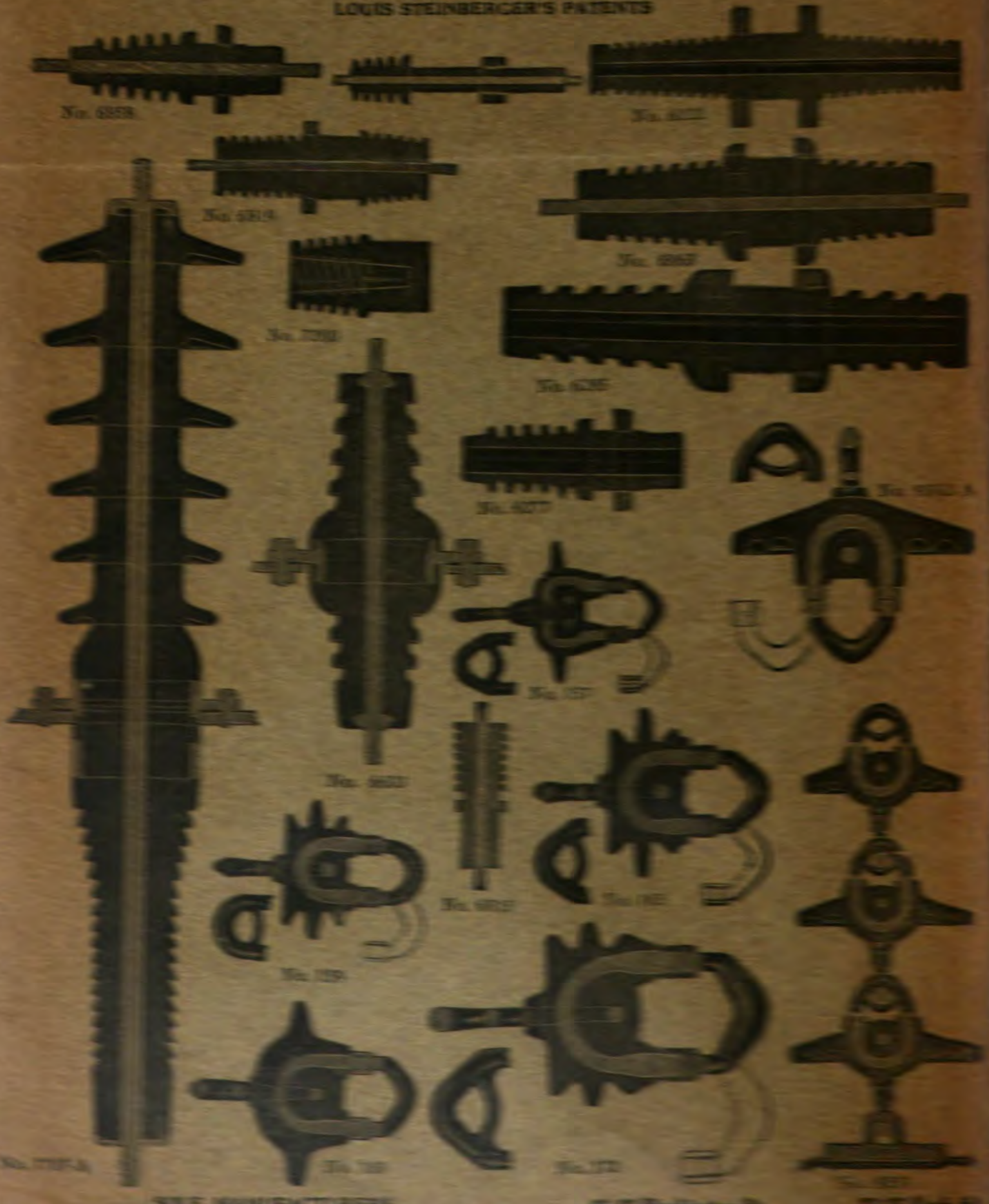

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